(51) International Patent Classification [6] : A61B 8/00	A1	(11) International Publication Number: WO 96/03919
		(43) International Publication Date: 15 February 1996 (15.02.96)

(54) Title: METHOD AND APPARATUS FOR A BASEBAND PROCESSOR OF A RECEIVE BEAMFORMER SYSTEM
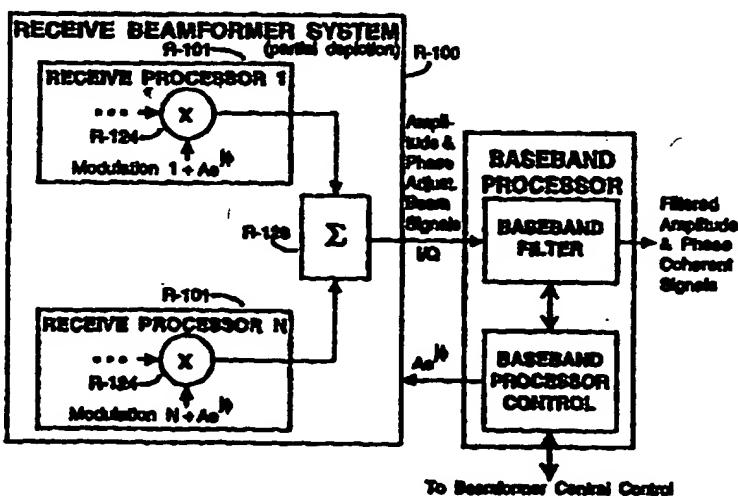
(57) Abstract

This invention presents a multi-beam base band processor (R-125) for making post-beam formation adjustments to the complex (imphase/quadrature) pre-detection scan line samples acquired from a receive beam former (R-100) of an ultrasound imaging system (R-20). Specifically, the invention includes a programmable finite impulse response (FIR) filter and a programmable complex multiplier. The programmable filter performs signal shaping and sample rate conversion. The signal shaping capability may be used to compensate for transducer and analog signal path responses, to increase signal to noise ratio (SNR) by rejecting out of band noise frequencies, to operate as a matched filter to the transmitted waveform shape, and to act as both a smoothing and anti-alias filter during rate conversion operation. The programmable multiplier permits both amplitude and phase adjustments of the complex data that assures amplitude and phase coherency between acquired scan lines, compensating for the effects of a selected scan format or the differences in ultrasound frequencies used to produce each scan line.

- 1 -

METHOD AND APPARATUS FOR A BASEBAND
PROCESSOR OF A RECEIVE BEAMFORMER SYSTEM

5

BACKGROUND OF THE INVENTION

This invention relates to coherent imaging systems
including, for example, radar, sonar, seismic, and
ultrasound systems, using vibratory energy, and in
10 particular, but not limited to, phased array ultrasound
imaging systems for scan formats such as, by way of
example only, linear, steered linear, sector, circular,
Vector®, steered Vector® in imaging modes such as, by way
of example only, B-mode (gray-scale) imaging mode and
15 color Doppler imaging mode. In particular, this
invention describes post-beamformation signal processing
adjustments. Although the invention will be discussed
with respect to an ultrasound system, the invention can
be implemented with other types of coherent imaging
20 systems.

Beam-to-beam coherency is not required in most prior
art ultrasonic imaging systems, although the need for
channel-to-channel coherency in phased array imaging
25 systems is well understood. In prior art systems, a
requirement for image uniformity is that the amplitude
response of the system to a point target at any range on
any scan line be substantially identical to the amplitude
response to the same target at the same range on an
30 adjacent scan line. The additional requirement for beam-
to-beam coherency further implies the phase response
(jointly represented with the amplitude response by, for
example, an in-phase and quadrature (I/Q) response) of
the system to a point target at any range on any scan
35 line also to be substantially identical to the phase
response to the same target at the same range on an
adjacent scan line. Systematic phase variations can

- 2 -

aris in some scan formats. For example, if the
apertures associated with successive transmit and receive
beams change relative to each other, systematic scan-
line-to-scan-line phase variations can be introduced.
5 Likewise, systematic scan-line-to-scan-line phase
variations can be introduced if the center frequencies of
successive transmit and receive beams change relative to
each other. This method requires range-dependent and
scan-line-dependent phase correction or adjustments of
10 such systematic variations. Such phase corrections or
adjustments may be predetermined and stored in memory to
be applied by a complex multiplier to the acquired
coherent samples prior to further coherent operations,
such as synthesizing new coherent samples.
15

It is also desirable to adjust for systematic phase
variations to establish coherent phase alignment among
pre-detected beams in a scan caused by differences in
beam-to-beam transmit/ receive frequencies by
20 remodulating prior to detection. This is most
efficiently performed on the beamformed baseband I/Q
signals.

Ultrasound imaging is accomplished by firing
25 (transmitting) into body tissue or other objects to be
imaged a scan sequence of focused ultrasonic beams
centered along straight lines in space called transmit
scan lines. The transmit scan lines are generated by a
transmit beamformer and an ultrasound transducer array.
30 The transmit scan lines are spaced to produce a planar
linear, planar sector or other display of the tissue via
a pre-defined firing or scanning pattern. Focused to
some defined depth in the tissue, the ultrasonic transmit
continuous-wave (CW) or pulse-wave (PW) signal,
35 propagating at an assum d constant propagation velocity
of nominally c=1540 m/sec through the tissue, interacts

with the tissue and reflects a small portion of the
signal back to the ultrasound transducer array that
initiated the ultrasound signal. The round trip delay
time is shortest for those targets closest to the
5 ultrasound transducer array, and longest for those
targets farthest from the transducer array. With the
application of appropriate time delays, the receive
beamformer can dynamically focus receive beams along
straight lines in space called receive scan lines
10 commencing, for example, with the shallowest range
(depth) of interest and evolving toward the deepest range
of interest.

Analog and hybrid (analog-digital) phased array
15 beamformer systems are available in the known art. For
example, phase array beamformer systems can be found in
the following patents.

U.S. Patent No.:	Title:	Inventor(s):
4,140,022	MULTIPLE TRANSDUCER ACOUSTIC IMAGING APPARATUS	Samuel H. Maslak

20

25

- 4 -

4,550,607	PHASED ARRAY ACOUSTIC IMAGING SYSTEM	Samuel H. Maslak J. Nelson Wright
5		
4,699,009	DYNAMICALLY FOCUSED LINEAR PHASED ARRAY ACOUSTIC IMAGING	Samuel H. Maslak Hugh G. Larsen
10	SYSTEM	
5,014,710 and 5,165,413	STEERED LINEAR COLOR DOPPLER IMAGING	Samuel H. Maslak Donald J. Burch J. Nelson Wright Hugh G. Larsen Donald R. Langdon Joel S. Chaffin Grant Flash, III

20

Digital receive beamformer systems have also been proposed in the art. The basic feature of a digital receive beamformer system can include: (1) amplification
25 of the ultrasound signal received at each element of an array such as, for example, a linear array; (2) direct per channel analog-to-digital conversion of the ultrasound signal with an analog-to-digital sampling rate at least twice the highest frequency in the signal; (3) a
30 digital memory to provide delays for focusing; and (4) digital summation of the focused signals from all the channels. Other processing features of a receive beamformer system can include phase rotation of a receive signal on a channel-by-channel basis to provide fine
35 focusing, amplitude scaling (apodization) to control the beam sidelobes, and digital filtering to control the bandwidth of the signal. This art points out the ever present desire to achieve, in an efficient manner, a reconstructed image of high quality.

40

- 5 -

SUMMARY OF THE INVENTION

The present invention relates to a method and
apparatus which provides for a baseband processor for
making post-beamformation adjustments to the complex (in-
5 phase/ quadrature) pre-detection scan line samples
acquired from a receive beamformer of an ultrasound
imaging system.

In a preferred embodiment, the invention includes a
10 programmable finite impulse response (FIR) filter and a
programmable complex multiplier. In one aspect of the
invention, the programmable filter performs signal
shaping. The signal shaping capability may be used to
compensate for transducer and analog signal path
15 responses that distort the transmitted and received
ultrasound waveforms. It may also be programmed to
increase signal-to-noise ratio (SNR) by rejecting out-of-
band noise frequencies. It may also be programmed to
operate as a matched filter to maximize the output SNR
20 relative to a selected transmit waveform shape. The
invention is capable of operating with time-interleaved
data samples associated with multiple beams.

In another aspect of the method and apparatus of the
25 invention, the baseband processor may operate as a sample
rate converter which converts the sample rate of the
receive beamformer signal to a rate that is advantageous
for an image display processor.

30   In another aspect of the method and apparatus of the
invention, the baseband processor includes a programmable
complex multiplier which follows the programmable filter.
The complex multiplier corrects systematic processing
amplitude and phase differences betwe n scan lines so
35 that there is scan-line-to-scan-lin (beam-to-beam)
amplitude coherence (also called amplitude matching) and

- 6 -

phase coherence (also called phase alignment). Such an aspect supports the novel feature of synthesizing samples and synthetic scan lines.

5   The method and apparatus further supports adjustable frequency scanning in order to mitigate grating lobes. The complex multiplier in another aspect can provide frequency remodulation of scan line data to maintain scan-line-to-scan-line phase coherence as the frequency

10 changes while scanning.

In still another aspect of this invention, the baseband processor can track receive signal frequency downshifting with tissue depth due to attenuation.

15

Additional advantages, objects, and novel features can be obtained from a review of the specification and the figures.

20   **BRIEF DESCRIPTION OF THE FIGURES**
Figs. 1a and 1b conceptually depict the transmission and reception of ultrasound beams to and from body tissue.

25   Fig. 2a depicts a block diagram schematic of a novel ultrasound beamformer system of an ultrasound medical imaging system including an embodiment of a baseband processor system of the invention.

30   Figs. 2b and 2c taken together depict a block diagram of the ultrasound beamformer system of Fig. 2a.

Fig. 3 depicts a block diagram of an embodiment of a digital baseband multi-beam processor for the receive

35 beamform r of Fig. 2.

- 7 -

Fig. 4a and 4b depict alternative embodiments of a baseband processor system. The first alternative in Fig. 4a switches the order of the baseband filter and complex multiplier. The second alternative in Fig. 4b moves the
5 amplitude and phase adjustment into the receive beamformer.

Fig. 5 is a block diagram of an embodiment of the baseband processor control of the invention which
10 provides for filtering and amplitude/phase adjustment among receive scan lines.

Figs. 6a, 6b and 6c depict graphs of typical signal frequency downshifting profiles that can be applied for
15 signal remodulation in the baseband processor.

Fig. 7 illustrates the waveform response at the same range point for three ultrasound scan lines.

20 Fig. 8 illustrates the method of phase accumulation at range r_t for a Vector scan geometry in which the phase differences for actual acquired scan lines (solid) at nominal line spacing are indicated. The dashed lines indicate target scan lines which were not acquired.
25

Fig. 9 demonstrates a plot (left) of the differential phase for a sequence of acquired ultrasonic scan lines (usl) and a plot (right) of the accumulated (integrated) phase.
30

DETAILED DESCRIPTION OF THE PRESENTLY PREFERRED EMBODIMENTS

35 The pr sent invention comprises a component of a medical ultrasound imaging system for which additional patent applications, listed above, have been

- 8 -

simultaneously filed in the United States Patent and
Trademark Office.

A.   Overview of Preferred Beamformer System Architecture

5

1.   Ultrasound Signal Description

Referring now to the drawings, Figs. 1a and 1b
depict representations of transmit and receive scan lines
10   (solid) and straight-line signal propagation paths from
individual elements (dashed), respectively. In Fig. 1a,
the transmit beamformer is generally identified by T-50
with the transducer array T-52 containing a multiplicity
of individual transducer elements T-54 organized as a
15   linear phased array in this particular embodiment. As is
known in the art, there are a great variety of transducer
array configurations available for use with ultrasound
transmit and receive beamformer systems. As can be seen
in Fig. 1a, the transmit beamformer T-50 sends
20   appropriately time-delayed electrical signals to the
individual transducer elements T-54. These transducer
elements T-54 then in turn convert electrical signals
into acoustic waves that propagate into the body tissue
T-56. By applying different time delays to the
25   excitation signals sent to the individual transducer
elements T-54, transmit scan lines T-60 and T-62, having
respective foci r_1 and r_2, can be established. It is to
be understood that each of these transmit scan lines is
representative of a center line of a different transmit
30   beam which is steered and focused into the body to be
imaged.

The transmit beamformer T-50 can generate
simultaneous multiple beams along different scan lines,
35   or different focal depths along the same scan line
(compound focus). Further, the multiple transmit beams
can each scan the entire image format or be transmitted

- 9 -

such that each of the multiple beams only scans a
specified section of the image format.

5   Fig. 1b depicts a digital receive beamformer R-58
which is also connected to the transducer array T-52.
Also depicted in Fig. 1b are receive scan lines R-64, R-
66 corresponding to a dynamically focused first receive
beam and a dynamically focused second receive beam,
respectively. The beams are sampled in range at a
10   plurality of focal depths (r_1, r_2, r_3) along each scan
line. In the digital receive signal path of the present
invention, transducer array signals can be selectively
separated into data representative of multiple individual
beams.
15

Each scan line of a transmit or receive scan pattern
can be parameterized by the origin on the transducer
array, the scan line orientation (angle θ) and the focus
depth or range (r). The ultrasound imaging system of the
20   present invention stores a pre-computed sparse data set
of focusing time delay and aperture apodization values
indexed by these parameters (based on geometric
considerations as is known in the art) and expands the
values by real-time computational means to control the
25   transmit and receive beamformation systems that produce
the desired scan lines.

2.   Beamformer System

Figs. 2a, 2b, 2c depict an overall block diagram of
30   a medical ultrasound imaging system R-20. Ultrasound
system R-20 includes a beamformer system R-22, one or
more transducers T-112, a display processing system R-26
with a display R-28 and an ultrasound imaging system
control R-40. As used herein, th t rm ultrasonic refers
35   to frequencies above the range of human hearing.

- 10 -

However, the transducer arrays T-112 are optimized for
frequencies typically within the range of 2-10 MHz.

In Figs. 2a, 2b, or 2c, the beamformer system R-22
5   includes inventive and novel (1) digital transmit
beamformer system T-102, (2) digital receive beamformer
system R-100, (3) beamformer central control system C-
104, (4) adaptive focusing control system G-100,
(5) Doppler receive beamformer system A-400, (6) baseband
10 multi-beam processor R-125, and (7) coherent sample
synthesizer S-100. These systems are depicted as
functional block diagrams. The blocks are abstracted
from the actual implementation of a preferred embodiment
in order to better illustrate the signal processing
15 functions performed.

In Figs. 2a, 2b, 2c, the control signals are
communicated over the light lead lines while the signal
paths are depicted with heavy lead lines.
20

As indicated in Figure 2a, beamformer system R-22
provides two sources of digital beam data to the display
processing system R-26:   (1) Doppler receive beamformer
single-beam complex in-phase/quadrature data representing
25 coherent temporal sampling of the beam (CW case) or
coherent temporal sampling at one range location along
the beam (PW case), and (2) digital receive beamformer
multi-beam complex in-phase/quadrature data representing
coherent sampling in range along each receive scan line.
30 Beamformer system R-22 can be operated to provide a
sequence of scan lines and associated samples as above to
provide data for a variety of display modes. By way of
example, possible display modes and their associated
processors include (1) brightness image and motion
35 processor R-30 for B-mode (gray-scale imaging) and M-mode
(motion display), (2) color Doppler image processor R-32

- 11 -

for flow imaging, and (3) spectral Doppler processor R-34
for wide dynamic nonimaging Doppl r velocity vs. time
displays. Additional display modes can be created from
the two complex data sources of R-22, as will be obvious
5 to those skilled in the art.

Ultrasound system R-20 also includes a transmit
demultiplexer T-106 for routing the output waveforms from
the transmitters T-103 to the transducer elements T-114,
10 a receive multiplexer R-108 for routing the input
waveforms from the transducer elements T-114 to the
receivers R-101, one or more transducer connectors T-110
and transducer arrays T-112. Many types of transducer
arrays can be used with the present system.
15

Ultrasound system R-20 also includes an ultrasound
imaging system control R-40, archival memory R-38 for
storing scan parameters and scan data, and operator
interface R-36.
20

The transducer array T-112 is interchangeable with a
variety of different kinds of transducer arrays,
including but not limited to linear, curved, curvi-linear
and annular transducer arrays. A variety of transducer
25 array shapes and frequencies are desirable in order to
satisfy the requirements of a variety of different
clinical settings. However, the transducer arrays T-112
are typically optimized for frequencies within the range
of 2-10 MHz. The medical ultrasound system R-20 performs
30 the three major functions of driving the ultrasonic
transducer array of elements T-114 to transmit focused
ultrasound energy, receiving and focusing back-scattered
ultrasound energy impinging on th transducer array T-
114, and controlling the transmit and receive functions
35 to scan a field of view in scan formats including (but
not limited to) linear, sector or Vector format.

- 12 -

3. Digital Transmit Beamformer System

The digital transmit beamformer T-102 (Fig. 2c) is
comprised of a plurality of digital multi-channel
transmitters T-103, one digital multi-channel

5   transmitters for one or more of the individual transducer
elements T-114. The transmitters are multi-channel in
that each transmitter can process, in a preferred
embodiment, up to four independent beams. Thus, for
example, 128 multi-channel transmitters have 512

10 channels. In other preferred embodiments, more than four
independent beams can be achieved. Processing more than
four beams per processor is within the scope of the
invention.

15   In a preferred embodiment, each of the digital
multi-channel transmitters T-103 produces as its output
in response to an excitation event the superposition of
up to four pulses, each pulse corresponding to a beam.
Each pulse has a precisely programmed waveform, whose

20 amplitude is apodized appropriately relative to the other
transmitters and/or channels, and delayed by a precisely
defined time delay relative to a common start-of-transmit
(SOT) signal. Transmitters T-103 are also capable of
producing CW.

25

Each digital multi-channel transmitter T-103
conceptually comprises a multiple beam transmit filter T-
115 which provides an output to a complex modulator T-
117. The output from complex modulator T-117 is

30 communicated to a delay/filter block T-119, and therefrom
is provided to a digital-to-analog converter (DAC) T-121.
The output of the DAC T-121 is amplified by an amplifier
T-123. The multiple beam transmit filter T-115, the
complex modulator T-117 and the delay/filter block T-119

35 comprise a digital multi-channel transmit processor T-
104.

- 13 -

The transmit filter T-115 can be programmed to
provide any arbitrary real or complex waveform responsive
to a start-of-transmit (SOT) signal. The transmit filter
T-115 is implemented with a memory which stores real or
5 complex samples of any desired and arbitrary pulse
waveform, and a means of reading the samples out
sequentially in response to the start-of-transmit (SOT)
signal delayed by a component of the focusing delay. In
a preferred embodiment, the memory of T-115 is programmed
10 to store baseband representations of real or complex
pulse envelopes.

Block T-115, although primarily a memory, is
referred to herein as a transmit filter, as the output of
15 block T-115 can be thought of as the time response of a
filter to an impulse. The complex modulator T-117
upconverts the envelope to the transmit frequency and
provides appropriate focusing phase and aperture
apodization.
20

Delay/filter block T-119 conceptually provides any
remaining focusing delay component and a final shaping
filter. The digital-to-analog converter (DAC) T-121
converts the transmit waveform samples to an analog
25 signal. The transmit amplifier T-123 sets the transmit
power level and generates the high-voltage signal which
is routed by the transmit demultiplexer T-106 to a
selected transducer element T-114.

30 Associated with each multi-channel transmit
processor T-104 is a local or secondary processor control
C-125 which provides control valu s and parameters, such
as apodization and delay values, to the functional blocks
of multi-chann l transmit proc ssor T-104. Each local or
35 secondary channel control C-125 is in turn controlled by
th central or primary control syst m C-104.

- 14 -

4. Digital Receive Beamformer System

The digital receive beamformer R-100 is illustrated
in Fig. 2b. The signals from the individual transducer
5 elements T-114 represent return echoes or return signals
which are reflected from the object being imaged. These
signals are communicated through the transducer
connectors T-110 to the receive multiplexer R-108.
Through multiplexer R-108, each transducer element T-114
10 is connected separately to one of the plurality of
digital multi-channel receivers R-101 which taken
together with summer R-126 comprise the digital receive
beamformer R-100 of the invention. The receivers are
multi-channel in that each receiver can process, in a
15 preferred embodiment, up to four independent beams.
Processing more than four beams per processor is within
the scope of the invention.

Each digital multi-channel receiver R-101 can, in a
20 preferred embodiment, comprise the following elements
which are represented by the function block diagram in
Fig. 2b. These elements include a dynamic low-noise and
variable time-gain amplifier R-116, an analog-to-digital
converter (ADC) R-118, and a digital multi-channel
25 receive processor R-120. The digital multi-channel
receive processor R-120 conceptually includes a
filter/delay unit R-122 and a complex demodulator R-124.
The filter/delay unit R-122 provides for filtering and
coarse focusing time delay. The complex demodulator R-
30 124 provides for fine focusing delay in the form of a
phase rotation and apodization (scaling or weighting), as
well as signal demodulation to or near baseband. The
digital multi-channel receivers R-101 communicate with
summer R-126 where the signal samples associated with
35 each beam from each receive processor are summed to form
final receive scan line samples, and the resulting

- 15 -

complex samples provided to baseband processor R-125.
The exact functioning and composition of each of these
blocks will be more fully described below.

5   A local or secondary control C-210 is associated
with each digital multi-channel receiver R-101. Local
processor control C-210 is controlled by central or
primary control C-104 and provides timing, control and
parameter values to each said receiver R-101. The
10   parameter values include focusing time delay profiles and
apodization profiles.

        5. Doppler Receive Beamformer System
        The Doppler receive beamformer system A-400 for wide
15   dynamic range, nonimaging Doppler acquisition includes
analog receivers A-402, each of which receives echo
signals from a respective one or more transducers T-114.
Each of the Doppler receivers A-402 includes a
demodulator/range gate A-404 which demodulates the
20   received signal and gates it (PW mode only) to select the
echo from a narrow range. The analog outputs of the
Doppler receivers A-402 are communicated to a Doppler
preprocessor A-406. In preprocessor A-406, the analog
signals are summed by summer A-408 and then integrated,
25   filtered, and sampled by analog processor A-410.
Preprocessor A-406 then digitizes the sampled analog
signal in an analog-to-digital converter (ADC) A-412.
The digitized signal is communicated to the display
processing system R-26.
30

        Associated with all Doppler receivers A-402 is a
single local or secondary Doppler beamformer control C-
127. Doppler beamformer control C-127 is controlled by
central or primary control system C-104 and provid s
35   control and focusing param ter values to th Doppler
receive beamformer system A-400.

- 16 -

The present beamformer system R-22 advantageously combines an imaging digital receive beamformation system R-100 and the nonimaging Doppler receive beamformation system A-400 in a manner which uses the same digital
5   transmit beamformation system T-102 and the same transducer array. This arrangement allows the digital receive beamformation system R-100 to be optimized for imaging modes such as B-mode and color Doppler imaging, and therefore has high spatial resolution, while the
10 accompanying Doppler receive beamformation system has a wide dynamic range and may be optimized for use in acquiring signals for nonimaging Doppler processing.

                6.   Beamformer Central Control System
15   The beamformer central control system C-104 of the present invention controls the operation of the digital transmit beamformer system T-102, the digital receive beamformer system R-100, the Doppler receive beamformer system A-400, the adaptive focusing control system G-100,
20 and the baseband processor R-125.

        The main control functions of the central control system C-104 are depicted in Fig. 2c. The control functions are implemented with four components. The
25 acquisition control C-130 communicates with the rest of the system including the ultrasound system control R-40 and provides high level control and downloading of scanning parameters. The focusing control C-132 computes in real time the dynamic delay and apodization digital
30 values required for transmit and receive beamformation, which includes pre-computed and expanded ideal values plus any estimated correction values provided by adaptive focusing control system G-100. The front end control C-134 sets the switches for the demultiplexer T-106 and the
35 multiplexer R-108, interfaces with the transducer connectors T-110, and sets the gain and bias levels of

- 17 -

all transmitter amplifiers T-123 and all receive
amplifiers R-116. The timing control C-136 provides all
the digital clocks required by the digital circuits.
This includes the sampling clocks for all the transmitter
5   DACs T-121 and receiver ADCs R-118.

        In a preferred embodiment central control C-104
expands sparse tables of focusing time delay and aperture
apodization values based on pre-computed and stored data,
10 through such techniques as interpolation and
extrapolation. The expanded delay and apodization values
are communicated to the local processor controls as a
profile of values across the transducer aperture where
the delay and apodization data expansion in range is
15 completed to per-transducer-element, per-sample, per-beam
values.

                7.   Adaptive Focusing Control System
        Adaptive focusing control system G-100 provides for
20 real time concurrent adaptive focusing. Adaptive
focusing control system G-100 is comprised of an adaptive
focus processor G-505 which provides focus correction
delay values to the focus control C-132 of the central
control C-104. Adaptive focus processor G-505 operates
25 on output produced by aberration value estimators G-502
from data gathered from the subarray summers R-126 of the
digital receive beamformer system R-100. Accordingly,
aberration correction values, preferably aberration delay
and amplitude values, are adaptively measured for each
30 receive scan line or for a subset of receive scan lines
in range regions corresponding to transmit focal depths
by the adaptive focusing control subsystem G-100 shown in
Figure 2c.

35   It is to be understood that in addition to the
adaptive focusing control syst m which adjusts focus

- 18 -

delays, that a number of adaptive control systems are contemplated. These systems, by way of example, include (1) adaptive contrast enhancement control system for adjusting focus delays and aperture apodizations, (2)

5   adaptive interference cancellation control for adjusting focus delays and phases, aperture apodizations, and (3) adaptive target enhancement control for adjusting focus delays and phase, aperture apodizations, imaging transmit and receive frequencies and baseband waveform shaping.

10

Another aspect of adaptive focusing which can be included in the preferred embodiment of the adaptive focusing control system G-100 is a geometric aberration transform device G-508/509 which can provide aberration

15   correction delay values to the adaptive focus processor G-505 for scan lines and scan line depth locations for which measured aberration values were not collected by aberration value estimators G-502. More specifically, measured aberration correction values are written to a

20   delay table in G-508/509. G-508/509 retrieves values from the delay table according to look-up rules of the geometric aberration transform to form focusing delay correction profiles across the aperture valid for depths, scan geometries, and acquisition modes other than the

25   depth, scan geometry, and mode for which aberration correction values were measured.

        8.   Baseband Processor System
        The baseband processor R-125 provides for filtering,
30   and receive-scan-line-to-receive-scan-line (beam-to-beam) amplitude and phase adjustments as discussed herein. The baseband processor R-125 additionally includes a baseband filter, a complex multiplier, and a bas band processor control which controls the operation of the baseband

35   filter and complex multiplier. The baseband processor control is controlled by central control C-104.

- 19 -

9. Coherent Sample Synthesizer System

The coherent sample synthesizer system S-100 is illustrated in Figure 2a. This system exploits the multi-beam transmit and multi-beam receive capability of

5 the invention to acquire and store coherent (pre-detection) samples of receive beam data along actual scan lines and to perform interpolation of the stored coherent samples to synthesize new coherent samples at new range locations along existing scan lines or along

10 synthetically-created scan lines. Both acquired and synthesized samples are passed to the display processing system R-26.

10. Transmit and Receive Multiplexers

15 The connectivity between the transducer array elements T-114 and the processors T-103, R-101, A-402 of the digital transmit, digital receive, and Doppler receive beamformer systems is established through a transmit demultiplexer T-106 and a separate receive

20 multiplexer R-108, as shown in Figure 2a. The multiple-transducer multiplexer configuration shown in Figure 2a permits selection of transmit and receive apertures lying entirely within a single transducer array or straddling across two transducer arrays. The two multiplexers are

25 independently controlled by the beamformer central control system C-104 and may be programmed to support a number of acquisition modes, including sliding aperture and synthetic aperture modes.

30 B. Baseband Processor System Preferred Embodiments

The digital baseband multi-beam processor R-125 (Fig. 3) comprises a baseband filter R-127, a complex multiplier B-254, and a baseband processor control C-270.

- 20 -

1. Baseband Filter and Complex Multiplier

The complex baseband beam signal (or time-
interleaved beam signals in the multiple beam case) from
the digital receive beamformer system R-100 is

5   communicated to the baseband multi-beam processor R-125.
The baseband filter R-127 performs filtering, signal
shaping, and/or sample rate conversion by a factor L/M,
where L is the integer interpolation factor and M is the
integer decimation factor. Complex multiplier B-254

10   provides for (1) scan-line-dependent and range-dependent
amplitude and phase adjustments of the filtered I/Q
signal B-260 required to correct for amplitude and phase
differences resulting from scan-line-to-scan-line
apodization changes, scan geometry, and non-aligned

15   effective transmit and receive origins, and (2)
remodulation (frequency adjustment) of the filtered
signal to correct for phase variations in range resulting
from different transmit frequencies per scan line. The
advantage of the use of a scan-line-to-scan-line variable

20   frequency mode on transmit and receive beamformation is
the reduction of grating lobes.

Such amplitude, phase, and frequency adjustments
between scan lines, particularly two or more adjacent

25   scan lines, is, for example, for purposes of creating
coherent samples needed to implement coherent image
processing techniques. Other post-beamformation
operations requiring phase-coherent samples between scan
lines will also benefit from the adjustments provided by

30   the complex multiplier.

Baseband filter R-127 preferably includes an
upsampler B-255 (Fig. 3), a downsampler B-257, and a
multi-tap FIR filter B-256, designated by the impuls

35   response sequence $h4[n]$, which is programmable using
either real or complex coefficients. The upsampler,

- 21 -

working in conjunction with the filter, acts as an
interpolator which increases the sample rate by an
upsampling integer factor L. The downsampler, working in
conjunction with the filter, acts as a decimator which
5   decreases the sample rate by a downsampling integer
factor M. When both the upsampler and downsampler are
operated, the net sample rate conversion factor is L/M.
The sample rate conversion is advantageous for adjusting
sample rates to downstream processors, such as the
10   display processing system R-26. In the preferred
embodiment, a 16-tap complex coefficient FIR filter will
satisfy the various filtering operations for which the
baseband filter may be programmed. During sample rate
conversion operation, fractionally-shifted versions of
15   the desired filter response in the coefficient registers
of the filter are appropriately sequenced during filter
operation.

Baseband filter R-127 can additionally accomplish
20   three other tasks. First, baseband filter R-127 can be
programmed to increase the signal-to-noise ratio by
rejecting out-of-band noise frequencies.

Second, the baseband filter may alternatively be
25   programmed to maximize the signal-to-noise ratio by
programming as a matched filter or quasi-matched filter
design, preferably for matching to substantially
Gaussian-shaped transmit pulses as well as pulses of
other shapes. Gaussian-shaped pulses are especially
30   useful as they represent waveforms that do not distort
during transmission through attenuative media such as the
body.

Third, baseband filt r R-127 enabl s pulse
35   equalization and shaping by compensating for both the

- 22 -

transducer frequency response and the analog signal path
prior to the ADC R-118.

2. Baseband Processor Control

5   The filter coefficients h4[n], decimation
factor M, and interpolation factor L for baseband filter .
R-127 are programmed into baseband processor R-125 by
being downloaded from the central control C-104 to
coefficient and rate memory C-278 (Fig. 5) in the
10   baseband processor control C-270 (Fig. 3). The
downloaded filter coefficients and sample rate conversion
factors can be changed at any time by introducing new
coefficients and factors into the central control C-104.
The coefficients and factors stored in the coefficient
15   and rate memory C-278 are selectable by the central
control C-104 for programming the filter and sample rate
conversion ratio L/M of the baseband filter R-127.

          Following complex multiplier B-254 is a register C-
20   296 which stores scan line sample data so that it can be
reported to the DMA processor C-202 of the central
control C-104 for providing scan-line-to-scan-line
calibration.

25   The complex multiplier B-254 includes a control
function which is contained in baseband processor control
C-270 (Fig. 5). In baseband processor control C-270, a
scan-line-to-scan-line (beam-to-beam) amplitude
adjustment value A and a phase adjustment value ϕ per
30   sample of each beam is generated in a time-interleaved
manner corresponding to the time-interleaved beam samples
B-260. As discussed above, the phase correction value ϕ
is the sum of phase terms that include: (1) a phase
adjustment term required to correct for phase differenc s
35   due to scan-line-to-scan-line apodization changes, (2) a
phase adjustment term for scan geometry which results in

- 23 -

non-aligned effective transmit and receive origins (the
scan-line-dependent and range-dependent phase adjustment
term), and (3) a phase term required to remodulate the
signal as though each scan line had been transmitted and
5   received at a common carrier frequency. Using a
frequency scaling factor or frequency vernier factor,
each beam can be produced with a different carrier
frequency. The complex multiplier accordingly provides
for remodulation of the received beam signals between
10 scan lines so that all beams are adjusted to account for
any differences in carrier frequencies.

    In operation a source data set including scan format
geometry parameters, sparse scan line gain and delay
15 values, interpolation coefficients for expanding the
sparse values, interpolation factor L, and decimation
factor M are downloaded from the central control C-104 to
the baseband processor control C-270. Additionally,
frequency parameters used in the frequency profile
20 generator of the central control C-104 in accordance with
Figs. 6a, 6b and 6c are downloaded to the baseband
processor control C-270. These profiles account for
frequency attenuation with depth.

25   The baseband processor control C-270 of Fig. 5
includes a gain and phase RAM C-280, a scan-line-indexed
interpolator C-282 which is supplied with pre-calculated
and pre-stored line interpolation coefficients (α_{line}) by
the central control C-104, and a range-indexed
30 interpolator C-284 with a range accumulator C-286, which
is supplied with a rational sample rate conversion factor
L/M and a phase zone width, both of which values are pre-
calculated and pre-stored in th central control C-104.
The rational sample rate conversion factor L/M is the
35 same L and M values supplied to th baseband filter R-
127. Additionally as is known in the art, a sample rate

- 24 -

conversion in accordance with the rational sample rate
conv rsion factor L/M is accomplished in order to match
the input sample data rate to the baseband filter R-127
with the output sample data rate to the display
5 processing system R-26.

Alternatively the range-indexed interpolator/
extrapolator C-284 can be supplied with programmable
interpolation/extrapolation coefficients which are, by
10 way of example, either (1) pre-calculated and pre-stored
in or calculated by the central control or (2) calculated
locally in baseband processor control C-270 by a
coefficient generator.

15 The baseband processor control C-270 also includes a
remodulation frequency processor C-292 which is
preferably implemented as a double phase accumulator.
The double phase accumulator calculates phase adjustment
values to correct for scan-line-to-scan-line frequency
20 differences and thus remodulates the signal as though a
common carrier frequency had been used across all scan
lines.

To understand the basis of the remodulation
25 operation, consider an idealized representation of a
signal at the output of a beamformer which has been
coherently summed across multiple elements, and has
undergone modulation on transmit, demodulation on
receive, and coherent summation:

$$x(t-2r/c) = e(t-2r/c) \cdot e^{j[\omega_m \cdot (t-2r/c)]} \cdot e^{-j[\omega_d \cdot t]} \qquad (1)$$

30

where,

e(t) = a baseband I/Q signal envelope,
$\omega_m = 2\pi f_m$ = a modulation frequency [MHz],
$\omega_d = 2\pi f_d$ = a demodulation frequency [MHz],
35 r = some imaging depth (range) [cm].

- 25 -

Note that the actual center frequency of the imaging
pulse, x(t-2r/c), depends additionally on other things,
such as tissue attenuation, filtering in the transmit and
receive processing chains, and other effects not
5   explicitly considered in the above equation (1). Also
not explicitly included in above equation (1) are the
detailed representations of focusing delay and phase
adjustments necessary for beamformation, though these
could be surmised by those skilled in the art. This
10   detail is not required to motivate the particular results
presented here.

The transmit modulation frequency, the receive
demodulation frequency, or both, may in general be range
15   dependent. In particular,

$$\omega_m = \omega_m(R_t), \text{ and } \omega_d = \omega_d(R_r),$$

where

R_t = the distance from the active array center to
the transmit focus,

20   R_r = the distance from the active array center to
the receive focus.

For a system with dynamic focus, this means that ω_d is
continuously updated.

25   We now consider a scan line 1 corresponding to a
modulation frequency ω_m^1, a demodulation frequency ω_d^1, and
a post-beamformer remodulation frequency ω_r^1; and an
adjacent scan line 2, with respective modulation,
demodulation, and remodulation frequencies ω_m^2, ω_d^2, ω_r^2.
30   It can be shown that the post-beamformed phase difference
between these two scan lines as a result of the different
modulation, demodulation, and remodulation frequencies
can be bounded by an amount $\Delta\nu$, wh re

$$\Delta\nu < (\omega_m^2 - \omega_m^1) \cdot T_p - [(\omega_d^2 + \omega_r^2) - (\omega_d^1 + \omega_r^1)] \cdot 2R_r/c \tag{2}$$

35   wh re,

- 26 -

T_p = the imaging pulse duration at any depth of the
receive beamformer signal output.

This expression is valid at the receive focal depth,
R_r, at the point of post-beamformer remodulation. It is
again noted that there may be other terms apart from $\Delta \nu$
which are needed to ensure phase coherence at the
beamformer output apart from the above equation (2).
Examples of such other terms include, but are not limited
to, terms which account for the offset in the beam
origin, such as naturally arise in Vector®, linear, and
curved linear formats, particularly with end alignment.
As expected, $\Delta \nu = 0$ when $\omega_m^2 = \omega_m^1$, $\omega_d^2 = \omega_d^1$, and $\omega_r^2 = \omega_r^1$.

We now make the observation, from the above equation
(2), that providing for remodulation at the post-
beamformer, pre-detected output with a frequency ω_r
permits scan-line-to-scan-line phase coherence by its
proper selection. In particular, by selecting ω_r^1 and ω_r^2
such that

$$\omega_d^1 + \omega_r^1 = \omega_d^2 + \omega_r^2 \qquad\qquad (3)$$

then the second term of equation (2) may be substantially
ignored. Note that if ω_d is range dependent, such as
would be the case for a range tracking system, then ω_r
must also be range dependent.

The first term of equation (2), given by $(\omega_m^2-\omega_m^1)\cdot T_p$
may be readily managed by keeping $(\omega_m^2-\omega_m^1)$ sufficiently
small. As an example, consider the requirement that
$\Delta \nu < \pi/4$, and suppose that, as might be typical, the
imaging pulse measured at the point of remodulation for a
tracking focused system has a duration that is four
cycles of the nominal modulation frequency. Then the
required limit on scan-line-to-scan-line frequency
variation becomes approximately, from equations (2) and
(3), $f_m^2 - f_m^1 < f_m^1/32$. If the nominal modulation

frequency is 5 Mhz, then the scan-line-to-scan-line
modulation frequency difference is constrained to be less
than 0.156 MHz, in this example.

5   Thus, if post-beamformation, pre-detection receive
processing requires beam-to-beam phase coherence for all
beams in a scan, then the maximum transmit carrier
frequency differential between any two beams in the scan
should be chosen to meet the above criteria.
10

        The above relationship (3) defining the remodulation
frequencies is independent of the modulation frequencies
on transmit. Such independence assumes that both the
modulation signal and the demodulation signal for all
15 transmit and receive channels are phase-locked to a
common timing clock reference. That is, the phases of
all such modulation and demodulation signals are defined
relative to a common time reference.

20   The above relationship (3) also assumes that the
modulation frequencies on successive transmit scan lines
and the demodulation frequencies on successive receive
scan lines are each slowly varying to avoid 2π phase
ambiguities. That is, $f_d^1 \approx f_d^2$ and $f_m^1 \approx f_m^2$. This
25 constraint is consistent with the problem being solved.

        The above relationship (3) also assumes a "well-
focused" system, wherein any observation made concerning
a point in the field of view occurs at a time when the
30 receive focus is at that point (i.e. tracking, or dynamic
focus), regardless of whether a target is also at that
point.

        From the central control C-104, pr -calculated and
35 pre-stored values representing the fr qu ncy differences
between scan lines (delta frequency values) are sent to

- 28 -

the remodulation frequency processor C-292. These
frequency difference values are based on frequencies and
frequency slopes such as specified in Figs. 6a, 6b and
6c. By way of example only, let it be assumed that the
5 frequency profiles for two scan lines look like Fig. 6b
but with different start frequency, F_{start}, values and
different downshift slope, $\Delta F_{downslope}$, values.
Accordingly, downloaded to baseband processor control C-
270 from the central control for the two scan lines are
10 the difference in frequencies between the scan lines and
the difference in the rate of change of the frequency
profiles over time. These values are calculated by the
acquisition processor C-130 based on stored parameters
and dependent upon the particular rational conversion
15 factor L/M currently being used. The first accumulator
of processor C-292 accumulates the difference in the
rates of change of the frequency profiles over time
between scan lines while the second accumulator
accumulates the difference in the frequencies between the
20 scan lines over time. If there is no difference in the
rate of change of the frequency profile over time, (i.e.
the profile are the same exact for initially different
F_{start} values, or after T_{break} in figure 6b when the slope
goes to zero) the first accumulator performs no function.
25 With no difference in the rate changes of the frequencies
between the scan lines, only the second accumulator
accumulates the frequency differences over time resulting
in a corrective remodulation phase value per sample.

30   The phase adjustment due to scan-line-to-scan-line
apodization changes, the phase adjustment due to scan
geometry which results in non-aligned transmit and
r ceive origins, and the phase adjustment due to
remodulating the signal to an effective common carrier
35 frequency are added in a summer C-288 and the summed
phase value is then converted in a look-up table C-290 to

- 29 -

sine and cosine representations. As part of the look-up
table C-290 function, the gain is multiplied by the sine
and cosine representations. This complex adjustment
value $Ae^{j\phi}$ is provided to complex multiplier B-254 for
5   application to the beam signals B-260.

It is to be understood that other embodiments of the
baseband processor control are possible within the scope
of this invention.
10

As indicated above, the complex multiplier B-254
ensures that coherent signal relationships are maintained
between scan lines. The transmit samples and the echo or
receive samples of the signals from beams are defined as
15   being coherent when sufficient information is stored,
preserved, or maintained to enable the samples of the
return signals to be phase and amplitude corrected from
scan-line-to-scan-line. The process of actually making
the phase and amplitude corrections need not have yet
20   taken place, as long as sufficient information with
respect to a reference is maintained.

When a signal sample is processed coherently, the
processing continues to maintain sufficient information
25   to perform phase and amplitude correction at a later
time. When two or more samples are processed coherently
(e.g., coherently summed), the phase and amplitude
corrections necessary to achieve phase and amplitude
coherence must have previously been performed.
30

Coherent processing of two or more signal samples
yields significant benefits, such as being able to
calculate synthetic samples, as described in the above
co-pending application.
35

- 30 -

Due to beamformer central control C-104 specifying
and accounting for all aspects of the transmit and
receive signal, the entire beamformer system maintains
all signal samples as coherent samples throughout the
5   transmit and receive signal path, until the signal is
finally detected in an operation which is external to
beamformation.

C.   Alternative Embodiments

10   In the preferred embodiment, the amplitude and phase
adjustments are accomplished by a complex multiplier
following the baseband filter. If x[n] represents a data
sample value to be adjusted, then $x[n] \cdot Ae^{j\phi}$ is the
adjusted value, where A is the amplitude adjustment value
15   and ϕ is the phase adjustment value. The same adjustment
can be accomplished by moving the complex multiplier in
the signal processing path to a position prior to the
baseband filter (Fig. 4a). Equivalently, the amplitude
and phase adjustment can be provided by moving the ...
20   operation into each of the receive processors R-101 of
the receive beamformer R-100 (Fig. 2b). In particular,
the demodulating value applied by each of the complex
demodulators R-124 to each receive processor channel
sample can be augmented by the amplitude and phase
25   adjustment factor, as illustrated in Figure 4b. Other
embodiments will be obvious to those skilled in the art.

D.   Basis for Amplitude and Phase Adjustments

The necessity for amplitude and phase correction,
30   applied after receive beamformation and baseband
filtering, can be understood by considering Fig. 7 which
shows the scan geometry of two acquired collinear B-mode
transmit/receive scan lines with a single target located
between these scan lines, at range r along a scan line
35   that passes through th targ t. The origins and angles
of the scan lin s are shown, with the dashed line shown

on each acquired transmit/receiv scan line pair
approximating the wavefront of the transmitt d wave at
range r_t, assuming the apodization used is symmetrical
about the scan line origins. The target scan lines,
5 although defined on the same grid as the acoustic
acquisition grid, need not coincide with one of the
acquired transmit/receive scan lines and, therefore, must
be synthetically created by, for example, interpolating
between two neighboring acquired transmit/receive scan
10 lines pairs.

        The interpolated scan line should appear as if it
were an actual scan line responding to the target, using
samples at range r_t from the corresponding neighboring
15 scan lines. However, the phase of the return signal at
range r_t on each acquired transmit/receive scan line is
not the same as the phase that would be measured at range
r_t had an actual scan line been acquired coinciding with
the target scan line. This is due to the slightly
20 different paths traveled by the waves emanating from each
acquired scan line, with origins essentially centered at
their apodization centroids, to the target and back.
Fig. 7 shows that the two-way path length for scan line 1
is greater than $2r_1$, while the path length for scan line 2
25 is less than $2r_1$. An additional component contributing to
a path length difference arises from the displacement of
the apodization centroid from the scan line origin, and
leads to additional phase error between the acquired
signals along scan line 1 and scan line 2. Therefore,
30 this phase difference between the scan line signals at
range r_t from each acquired scan line exists and must be
compensated before coherent processing such as
interpolation betw n scan lines is performed. This is
illustrated by the return signal waveforms of Fig. 7.
35 Suppose that th phase measur d at rang r_t along scan
        line 2 differs by 180° from the phase measured at range r_t

- 32 -

along scan line 1, given a target centered between scan
line 1 and scan line 2 at range r_t. If the amplitudes of
the return signals were approximately equal, then
interpolation between scan line 1 and scan line 2 at
5   range r_t would yield essentially a zero result. However,
if an actual collinear transmit/receive scan line pair   .
were fired along the target scan line, the non-zero
response shown by the dashed waveform in Fig. 7 would be
the actual result. In general, the phase difference to
10   be compensated by the baseband processor is a function of
range and scan line origin and angle.

     In general, there are both amplitude and phase
differences between acquired scan line signals at a range
15   r, given a target range r along a target scan line, which
depend in a complicated relationship on the scan format,
array shape, target geometry with respect to the array,
transmit/receive apodizations, transmit/receive
apodization end-treatment, single firing or multiple
20   firing synthetic scan line formation, frequency, and
waveform shape. An approximate closed form expression
for the scan line relative response to a reference point
amplitude and phase is based on the assumptions that (1)
target scan line lies within the main beam of each of the
25   respective scan line pairs, i.e., the transmit and
receive scan lines are spaced less than or equal to the
Nyquist spacing, and (2) the apodization function is a
Gaussian weighting across the array aperture. The
relative response at range r_t to an absolute reference
30   phasor ($Ae^{j\phi}$) is another complex phasor that is a function
of three major terms. The relative response phasor $\rho(r_t)$
is given by the set of equations in Table 1. Note the
list of dependent parameters at the end of Table 1, which
include all typical scanning format parameters
35   ncountered in ultrasonic imaging.

- 33 -

To compensate for the effect of this difference, complex beam samples could, for example, simply be multiplied by the inverse $1/\rho(r_t)$ before coherent operations on samples at a common range, but different

5   scan lines, is performed. Alternately, only the relative phasor difference between two spatial points in the acquired scan line data needs to be corrected before coherent operations are performed with the data. Assuming a uniform target has been intersected by the

10 scan lines, let $\rho_1 = A_1 \cdot e^{j\Psi_1}$ be the relative (to a reference point) complex phasor response at a point 1 along some scan line, and let $\rho_2 = A_2 \cdot e^{j\Psi_2}$ be the relative complex phasor response at another point 2 elsewhere (could be the on the same scan line or on another scan

15 line). Ideally, one would like the signal processing performed by the beamformation to yield identical complex beam signal responses everywhere that the uniform target is intersected by the scan lines so that, for example, linear interpolation between any two scan lines samples

20 would yield the identical complex response. When such a condition is achieved, the scan line signal responses are said to be "amplitude coherent" (because $A_1 = A_2$ at the two points) and "phase coherent" (because $\Psi_1 = \Psi_2$ at the two points). However, geometric conditions of

25 acquisition and the finiteness of apertures produces systematic effects that lead to different complex responses, as Table 1 clearly indicates. One must at least adjust the "relative" response at two points, using for example "complex multiplication" of the point 1

30 response by ρ_2/ρ_1 or "complex multiplication" of the point 2 response by ρ_1/ρ_2, before there is amplitude coherence and phase coherence between at least these two spatial locations. Becaus $\rho_2/\rho_1 = (A_2/A_1)e^{j(\Psi_2 - \Psi_1)}$ or $\rho_1/\rho_2 = (A_1/A_2)e^{j(\Psi_1 - \Psi_2)}$, amplitude coher nc is achieved by

35 "amplitude adjustment" using the factor A_2/A_1 or A_1/A_2, and phase coherence is achi ved by "phase adjustment" by the

- 34 -

factor $(\Psi_2 - \Psi_1)$ or $(\Psi_1 - \Psi_2)$. The process of achieving
phase coherence is also called "phase alignment" because
the relative phase between two spatial sample points
after phase adjustment is such that $(\Psi_2 - \Psi_1) = 0$, that
5   is, the phase responses are "aligned" after adjustment to
yield a zero phase difference. The term complex response
adjustment will also be used here to mean that one or
both phase adjustment and amplitude adjustment operations
are to be used, unless indicated otherwise.

10   Because a complex number c may be represented in
polar coordinate form as $c = a \cdot e^{j\theta}$ or equivalently as the
in-phase and quadrature (I/Q) format $c = x_I + j x_Q$, in which
$a = (x_I^2 + x_Q^2)^{1/2}$, $\theta = \arctan (x_I/x_Q)$, $x_I = a \cdot \cos\theta$, $x_Q = a \cdot \sin\theta$, the
amplitude and phase adjustments to yield $\rho \cdot c$ using some
15   complex value ρ (or ratio of ρ values) may be performed
using a number of means. A "complex multiplier" is the
most generic means for applying a complex adjustment
factor ρ to one or more complex beam signal samples. If
the polar format representation for ρ is used, then
20   separate steps using a single "real multiplication" by
the amplitude factor "a" and using a "phase rotation" by
a phase factor θ are the appropriate enabling means. If
the in-phase and quadrature format representation for ρ
is used, then separate "real multiplications" by the two
25   factors $a \cdot \cos\theta$ and $a \cdot \sin\theta$ are the appropriate means, in
which case the θ factor is represented as a "phase shift"
operation. Describing a "complex multiplier" here
includes those arrangements involving or using real
multipliers, phase rotators, and phase shifters.

30   The three term product of the phasor $\rho(r_t)$ has a
geometric term G_ρ, an apodization term A_ρ, and an origin
term O_ρ. G_ρ is dominated by dependencies on the
transducer array and scan geometry and includes broadband
pulse effects. Observe that for targets at increasing
35   ranges from th transmit/receive scan line axis, term
$(1 + \sigma^2 \beta^2)$ increases so that the two-way Gaussian pulse

- 35 -

bandwidth and c nter fr quency decreases due to the
lowpass filtering effect of temporal dispersion. The
term A_p is dominated by dependencies on the apodization
which is assumed to have Gaussian shape. The origin term
5   O_p is simply the product of the transmit and receive
origin terms O_x and O_r and contains the beam loss and an
additional phase shift term imposed on the beam due to
its transmit/receive apodization origins x_{ax} and x_{ar} being
displaced from the transmit/receive scan line origins x_{ax}
10   and x_{ar} respectively. O_p also accounts for the effective
apodization shifting due to element factor and
attenuation effects.

Denoting the relative phasor response at range r_t
15   along scan line 1 as $\rho_1(r_t)=A_1e^{j\psi_1}$ and along scan line 2 as
$\rho_2(r_t)=A_2e^{j\psi_2}$, where A_1 and A_2 are the relative amplitude
responses and where ψ_1 and ψ_2 are the relative phase
responses, the amplitude difference that requires
amplitude adjustment of one or both scan line responses
20   at range r_t in order to match amplitude response involves
the gain factor A_2/A_1 (or A_1/A_2), and the phase difference
that requires phase adjustment of one or both scan line
responses at range r_t is the factor $\psi_1-\psi_2$ (or $\psi_2-\psi_1$).
There is one scan format that basically requires no
25   amplitude or phase adjustments in order to maintain
amplitude and phase coherence for purposes, as an
example, of interpolating new scan line samples.
Consider a sector scan format of a linear array in which
the scan vertex is located at the center of the array.
30   Also assume that the same imaging center frequency is
used to produce all scan lines and a symmetric
apodization is applied. In this case,
$x_{ox} = x_{or} = x_t = x_{ax} = x_{ar}$ and the various differenc
expressions, such as $x_{tox} = x_t - x_{ox}$, are all z ro. In this
35   situation, the equation in Table 1 yi lds phase terms
(having the form e^j for some phase argument ψ) which are

- 36 -

essentially zero, and amplitude terms for a given range
that are id ntical for all scan lines. Thus, the sector
scan format generally requires no amplitude or phase
adjustments as there is already nearly full coherence
5   among scan lines at a given range.

Although the phasor equation of Table 1 provides an
analytical expression for a systematic understanding of
amplitude and phase adjustments needed to maintain inter-
10   scan-line coherence, evaluating such an expression as
part of a real-time scanning operation would be
prohibitive. An alternative approach amenable to real-
time implementation is to precalculate amplitude and
phase correction values evaluated on a sparse grid of
15   reference ranges indexed by scan line number.
Interpolation into the sparse grid of amplitude and phase
correction values for arbitrary range and/or arbitrary
scan lines between indexed scan lines can then produce
approximate amplitude and phase corrections that are
20   applied by the complex multiplier B-246, or comparable
apparatus such as real multiplier for amplitude and phase
rotator for phase, to each sample of each scan line.

Consider a specific case of generating the phase
25   adjustment. The phase portion of the correction is
simply the difference of the individual scan line
relative phase responses.

$$\psi_2(r_t) - \psi_1(r_t) = -\Delta\phi(r_t)$$

Although the phase difference $\Delta\phi$ can be computed across
30   scan lines, it does not determine the individual scan
line phases ψ_1 and ψ_2 needed to phase correct or phase
align the individual scan lin samples. Consider Fig. 8
which shows the phase differences for a group of scan
lines spac d at a nominal scan line spacing, where the
35   dashed lines indicate target scan lines. The phas ψ for

- 37 -

a given phase correction reference scan line, which for
single beam scanning modes is set equal to the target
scan line, is determined such that the phase differences
formed from neighboring ψ result in either using $-\Delta\phi$
5   directly, or is the result of a linear interpolation of
neighboring phase differences. Starting from the center
scan line and scanning to the right in Fig. 8, the phases
are accumulated in the following manner. The phase ψ_1 is
set to zero since the absolute phase reference is
10   completely arbitrary; $\psi_2 = -1/2 \cdot (\Delta\phi_{21} + \Delta\phi_{32})$,

$\psi_3 = \psi_2 - 1/2 \cdot (\Delta\phi_{32} + \Delta\phi_{43})$, etc. This is simply numerical
         of integral differential phases across reference
scan lines, which in this discussion employs the
trapezoidal rule. Trapezoidal integration approximates
15   the integral of a function by accumulating the areas of
trapezoids. This is shown in Fig. 9 where the dashed
line indicate the trapezoidal approximation to the
continuous differential phase correction values indicated
by the solid line. The resulting trapezoidal
20   integration is given mathematically by

$\psi_1 = 0$

$$\psi_2 = \psi_1 - \frac{1}{2} (\Delta\phi_{21} + \Delta\phi_{32}) = -\frac{1}{2} (\Delta\phi_{21} + \Delta\phi_{32})$$

$$\psi_3 = \psi_2 - \frac{1}{2} (\Delta\phi_{32} + \Delta\phi_{43}) = -\frac{1}{2} (\Delta\phi_{21} + 2\Delta\phi_{32} + \Delta\phi_{43})$$

$$\psi_4 = \psi_3 - \frac{1}{2} (\Delta\phi_{43} + \Delta\phi_{54}) = -\frac{1}{2} (\Delta\phi_{21} + 2\Delta\phi_{32} + 2\Delta\phi_{43} + \Delta\phi_{54})$$

25

and so forth.

During real-time B-mode imaging, the baseband
processor will apply the range-varying accumulated phase
correction values to each range sample of each receive
30   scan line. If a given range sample does not coincide
with a phase correction value in the precalculated table
at a reference scan line and a reference range, the

- 38 -

baseband processor control linearly interpolates among
the accumulated phase correction values in scan line and
range to produce a correction phase. Practical
consideration of memory size will limit the number of
5   reference ranges and reference scan lines, so it is
desirable to distribute the reference points in the
range-scan-line grid in a manner which minimizes the
phase error between desired phase correction values and
interpolated phase correction values over the image
10   region. The differential phase varies most rapidly near
the transmit focus. Since the phase accumulation occurs
across scan lines, the phase correction values will also
vary most rapidly near the transmit focus. Therefore, it
is advantageous to place a higher density of reference
15   ranges near the transmit focus to accurately track phase
variations in range by the piecewise linear approximation
approach which has been described, and a lower density of
reference ranges in regions which exhibit phase
variations which vary slowly with range. The
20   distribution of reference scan lines will therefore use
non-uniform spacing, with the highest concentration of
reference ranges where the phase changes rapidly.

TABLE 1

$$\rho(r_t) = G_\rho \cdot A_\rho \cdot O_\rho$$

$$\rho(r_t) = \underbrace{G_0 \, e^{-\pi \frac{\sigma^2 (2r_t - R_{to})^2}{c^2 (1 + \sigma^2 \beta^2)}} e^{\frac{2\pi j f_c(r_t)(2r_t - R_{to})}{c(1 + \sigma^2 \beta^2)}}}_{G_\rho} \underbrace{\left[\frac{W_{1tx}}{\alpha_x(\lambda_c)}\right]\left[\frac{W_{1tr}}{\alpha_r(\lambda_c)}\right] \frac{e^{-\pi \frac{\beta^2 f_c^2(r_t)}{(1 + \sigma^2 \beta^2)}}}{\sqrt{1 + \sigma^2 \beta^2}}}_{A_\rho} \underbrace{O_x(\lambda_c) O_r(\lambda_c)}_{O_\rho}$$

$$f_c(r_t) = \frac{\sigma^2}{\sigma_{sys}^2}\left[f_{c0} - \frac{\gamma}{2\pi}\sigma_{sys}^2 (R_{tax} + R_{tar})\right] + \frac{\sigma^2}{\sigma_{BBF}^2} f_d(r_t) \qquad \lambda_c(r_t) = \frac{c}{f_c(r_t)} \qquad \sigma = \frac{\sigma_{sys}\,\sigma_{BBF}}{\sqrt{\sigma_{sys}^2 + \sigma_{BBF}^2}}$$

$$R_{to} = R_{tox} + R_{tor} = 2r_t + \sin\theta_{t\perp}(x_{tox} + x_{tor}) + \frac{1}{2}\left(\frac{\cos^2\theta_{t\perp}}{r_t} + \frac{\cos\theta_{t\perp}}{R_{xdcr}}\right)(x_{tox}^2 + x_{tor}^2)$$

$$R_{tax} = \sqrt{r_t^2 - 2r_t \sin\theta_{t\perp}\, x_{atx} + \left(1 + \frac{r_t}{R_{xdcr}}\cos\theta_{t\perp}\right)x_{atx}^2}$$

$$R_{tar} = \sqrt{r_t^2 - 2r_t \sin\theta_{t\perp}\, x_{atr} + \left(1 + \frac{r_t}{R_{xdcr}}\cos\theta_{t\perp}\right)x_{atr}^2}$$

$$\beta^2 = \beta_x^2(\lambda_c) + \beta_{ex}^2 + \beta_r^2(\lambda_c) + \beta_{er}^2 \qquad G_0 = \frac{e^{\pi\left[\frac{f_c^2(r_t)}{\sigma^2} - \frac{f_{c0}^2}{\sigma_{sys}^2} - \frac{f_d^2(r_t)}{\sigma_{BBF}^2}\right]}}{16\pi^2 R_{tax} R_{tar} \sqrt{\sigma_{sys}^2 + \sigma_{BBF}^2}}$$

$$\beta_x^2(\lambda_c) = \frac{W_{1tx}^2 s_x^2}{\alpha_x^2(\lambda_c) c^2} \qquad s_x = (\sin\theta_{t\perp} - \sin\theta_{x\perp}) + \left(\frac{\cos^2\theta_{t\perp}}{r_t} + \frac{\cos\theta_{t\perp}}{R_{xdcr}}\right)x_{tox}$$

$$\beta_r^2(\lambda_c) = \frac{W_{1tr}^2 s_r^2}{\alpha_r^2(\lambda_c) c^2} \qquad s_r = (\sin\theta_{t\perp} - \sin\theta_{r\perp}) + \left(\frac{\cos^2\theta_{t\perp}}{r_t} + \frac{\cos\theta_{t\perp}}{R_{xdcr}}\right)x_{tor}$$

$$\beta_{ex}^2 = \frac{W_e^2 \sin^2\theta_{eax}}{c^2} \qquad \sin\theta_{eax} = \frac{\left[r_t \sin\theta_{t\perp} - \left(1 + \frac{r_t}{R_{xdcr}}\cos\theta_{t\perp}\right)x_{atx}\right]}{R_{tax}}$$

$$\beta_{er}^2 = \frac{W_e^2 \sin^2\theta_{ear}}{c^2} \qquad \sin\theta_{ear} = \frac{\left[r_t \sin\theta_{t\perp} - \left(1 + \frac{r_t}{R_{xdcr}}\cos\theta_{t\perp}\right)x_{atr}\right]}{R_{tar}}$$

$$\alpha_x^2(\lambda_c) = \left\{1 + j\frac{W_{1tx}^2}{\lambda_c}\left[\left(\frac{\cos^2\theta_{t\perp}}{r_t} + \frac{\cos\theta_{t\perp}}{R_{xdcr}}\right) - \left(\frac{\cos^2\theta_{x\perp}}{r_x} + \frac{\cos\theta_{x\perp}}{R_{xdcr}}\right)\right]\right\}$$

$$\alpha_r^2(\lambda_c) = \left\{1 + j\frac{W_{1tr}^2}{\lambda_c}\left[\left(\frac{\cos^2\theta_{t\perp}}{r_t} + \frac{\cos\theta_{t\perp}}{R_{xdcr}}\right) - \left(\frac{\cos^2\theta_{r\perp}}{r_r} + \frac{\cos\theta_{r\perp}}{R_{xdcr}}\right)\right]\right\}$$

- 40 -

$$O_x(\lambda_c) = E_{0x} F_{0x} e^{\frac{2\pi j s_x x_{1 o x}}{\lambda_c \alpha_c^2(\lambda)}} e^{-\pi \frac{\left[1 - \frac{1}{\alpha_c^2(\lambda_c)}\right] x_{1 o x}}{W_{1 l x}^2}} \qquad O_r(\lambda_c) = E_{0r} F_{0r} e^{\frac{2\pi j s_x x_{1 o r}}{\lambda_c \alpha_c^2(\lambda)}} e^{-\pi \frac{\left[1 - \frac{1}{\alpha_c^2(\lambda_c)}\right] x_{1 o r}}{W_{1 l r}^2}}$$

$$E_{0x} = W_e \cos \theta_{eax} \operatorname{sinc}\left(\frac{W_e \sin \theta_{eax}}{\lambda_c}\right) e^{\pi \frac{W_e^2 \sin^2 \theta_{eax}}{\lambda^2}} e^{\pi \frac{(x_{eox} - x_{efox})^2}{W_{efx}^2}}$$

$$E_{0r} = W_e \cos \theta_{ear} \operatorname{sinc}\left(\frac{W_e \sin \theta_{ear}}{\lambda_c}\right) e^{\pi \frac{W_e^2 \sin^2 \theta_{ear}}{\lambda^2}} e^{\pi \frac{(x_{eor} - x_{efor})^2}{W_{efr}^2}}$$

$$\cos \theta_{eax} = \frac{r_t \cos \theta_{t\perp} + \frac{r_t}{R_{xdcr}} \sin \theta_{t\perp} x_{atx} - \frac{1}{2}\left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp}\right)\frac{x_{atx}^2}{R_{xdcr}}}{R_{tax}}$$

$$\cos \theta_{ear} = \frac{r_t \cos \theta_{t\perp} + \frac{r_t}{R_{xdcr}} \sin \theta_{t\perp} x_{atr} - \frac{1}{2}\left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp}\right)\frac{x_{atr}^2}{R_{xdcr}}}{R_{tar}}$$

$$F_{0x} = e^{\pi \frac{(x_{eox} - x_{\gamma ox})^2}{W_{\gamma x}^2}} e^{\pi \frac{x_{1 o x}}{W_{1 l x}^2}} e^{-\pi\left(\frac{x_{eox}^2}{W_x^2} + \frac{x_{efox}^2}{W_{efx}^2} + \frac{x_{\gamma ox}^2}{W_{\gamma x}^2}\right)}$$

$$F_{0r} = e^{\pi \frac{(x_{eor} - x_{\gamma or})^2}{W_{\gamma r}^2}} e^{\pi \frac{x_{1 o r}}{W_{1 l r}^2}} e^{-\pi\left(\frac{x_{eor}^2}{W_r^2} + \frac{x_{efor}^2}{W_{efr}^2} + \frac{x_{\gamma or}^2}{W_{\gamma r}^2}\right)}$$

$$x_{1 l o x} = W_{1 l x}^2\left(\frac{x_{aox}}{W_x^2} + \frac{x_{efox}}{W_{efx}^2} + \frac{x_{\gamma o x}}{W_{\gamma x}^2}\right) \qquad W_{1 l x} = \frac{1}{\sqrt{\frac{1}{W_x^2} + \frac{1}{W_{efx}^2} + \frac{1}{W_{\gamma x}^2}}}$$

$$x_{1 l o r} = W_{1 l r}^2\left(\frac{x_{aor}}{W_r^2} + \frac{x_{efor}}{W_{efr}^2} + \frac{x_{\gamma o r}}{W_{\gamma r}^2}\right) \qquad W_{1 l r} = \frac{1}{\sqrt{\frac{1}{W_r^2} + \frac{1}{W_{efr}^2} + \frac{1}{W_{\gamma r}^2}}}$$

$$x_{efox} = x_{tox} + \frac{r_t \sin \theta_{t\perp}}{\left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp}\right)} \qquad W_{efx} = \frac{\lambda_c R_{tax}}{W_e\left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp}\right)}$$

$$x_{efor} = x_{tor} + \frac{r_t \sin \theta_{t\perp}}{\left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp}\right)} \qquad W_{efr} = \frac{\lambda_c R_{tar}}{W_e\left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp}\right)}$$

- 41 -

$$x_{\gamma ox} = x_{aox} + \frac{R_{tax}^2 \left[r_t \sin \theta_{t\perp} - \left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp} \right) x_{atx} \right]}{r_t^2 \left(\cos^2 \theta_{t\perp} + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp} \right)}$$

$$W_{\gamma x} = \sqrt{\frac{2\pi R_{tax}^3}{\gamma f_c \, r_t^2 \left(\cos^2 \theta_{t\perp} + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp} \right)}}$$

$$x_{\gamma or} = x_{aor} + \frac{R_{tar}^2 \left[r_t \sin \theta_{t\perp} - \left(1 + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp} \right) x_{atr} \right]}{r_t^2 \left(\cos^2 \theta_{t\perp} + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp} \right)}$$

$$W_{\gamma r} = \sqrt{\frac{2\pi R_{tar}^3}{\gamma f_c \, r_t^2 \left(\cos^2 \theta_{t\perp} + \frac{r_t}{R_{xdcr}} \cos \theta_{t\perp} \right)}}$$

$$x_{ax} = \frac{\int_{-\infty}^{\infty} a_x(x) \, x \, dx}{W_x} \approx \frac{d \cdot \sum_{n=0}^{N-1} a_x(x_n) \, x_n}{W_x} \qquad W_x = \int_{-\infty}^{\infty} a_x(x) \, dx \approx d \cdot \sum_{n=0}^{N-1} a_x(x_n)$$

$$x_{ar} = \frac{\int_{-\infty}^{\infty} a_r(x) \, x \, dx}{W_r} \approx \frac{d \cdot \sum_{n=0}^{N-1} a_r(x_n) \, x_n}{W_r} \qquad W_r = \int_{-\infty}^{\infty} a_r(x) \, dx \approx d \cdot \sum_{n=0}^{N-1} a_r(x_n)$$

$$x_{tox} = x_t - x_{ox} \qquad\qquad x_{aox} = x_{ax} - x_{ox} \qquad\qquad x_{atx} = x_{ax} - x_t$$

$$x_{tor} = x_t - x_{or} \qquad\qquad x_{aor} = x_{ar} - x_{or} \qquad\qquad x_{atr} = x_{ar} - x_t$$

$f_c(r_t)$ = effective imaging center frequency f_{c0} = imaging center frequency (zero depth)

$f_d(r_t)$ = demodulation frequency $\psi_d(r_t)$ = demodulation phase γ = attenuation

r_x = transmit focus range r_r = receive focus range r_t = target range

$\theta_{x\perp}$ = transmit usl angle $\theta_{r\perp}$ = receive usl angle $\theta_{t\perp}$ = target usl angle

x_{ox} = transmit usl origin x_{or} = receive usl origin x_t = target usl origin

x_{ax} = transmit gaussian apodization origin x_{ar} = receive gaussian apodization origin

W_x = transmit gaussian apodization width W_r = receive gaussian apodization width

d = element spacing W_e = element width R_{xdcr} = transducer radius N = # of elements

Note: Angles are with respect to th array surface normal and c is the acoustic propagation velocity.

- 42 -

WE CLAIM:

1.   A baseband processor for processing signals
from an ultrasound receive beamformer which generates
5   samples representative of an ultrasound signal,
comprising:
       a complex response adjuster, said complex
response adjuster operating on said samples to effect
coherence between samples.
10

2.   The baseband processor of claim 1 wherein the
complex response adjuster comprises a phase adjuster,
said phase adjuster operating on said beamformed samples
to effect phase coherence between said samples, thereby
15   creating phase coherent samples.

3.   The baseband processor of claim 2 further
comprising an amplitude adjuster which operates on the
beamformed samples to effect amplitude coherence between
20   said samples, thereby creating amplitude coherent
samples.

4.   The baseband processor of claim 2 wherein:
       said phase adjuster operates on the phase of
25   the beamformed samples to effect phase coherence between
samples on different scan lines.

5.   The baseband processor of claim 2 wherein:
       said phase adjuster operates on the beamformed
30   samples to effect phase coherence between samples on the
same scan line.

6.   The baseband processor of claim 2 wherein:
       said phase adjuster remodulates the beamformed
35   samples on a receive scan line.

- 43 -

7. The baseband processor of claim 6 wherein said phase adjuster remodulates the beamformed samples associated with distinct beams using different frequencies to effect beam-to-beam coherence.

5

8. The baseband processor of claim 2 wherein the phase adjuster comprises a phase rotator.

9. The baseband processor of claim 2 wherein the phase adjuster comprises a complex value multiplier.

10. The baseband processor of claim 1 further comprising a baseband filter.

11. The baseband processor of claim 10 wherein the baseband filter is connected to receive as inputs said beamformed samples and wherein said complex response adjuster is connected to receive the output of the baseband filter.

12. The baseband processor of claim 10 wherein the complex response adjuster is connected to receive as inputs said beamformed samples and said baseband filter is connected to receive the output of said complex response adjuster.

13. The baseband processor of claim 10 wherein the baseband filter comprises a programmable filter.

14. The baseband processor of claim 10 or 13 wherein the baseband filter comprises a sample rate converter.

15. The baseband processor of claim 10 wherein:

- 44 -

said baseband filter increases the signal to noise ratio of said signals from said ultrasound beamformer by rejecting out of band noise frequencies.

5   16.   The baseband processor of claim 10 wherein said baseband filter is adapted to shape said signals from said ultrasound beamformer.

        17.   The baseband processor of claim 16 wherein:
10   said baseband filter is adapted to produce an approximately Gaussian-shaped waveform.

        18.   The baseband processor of claim 1 or 10 wherein the complex response adjuster comprises:
15   (a)   a phase adjuster, said phase adjuster operating on said samples to create phase coherent samples; and
              (b)   an amplitude adjuster, said amplitude adjuster operating on said samples to create amplitude
20   coherent samples.

        19.   The baseband processor of claim 1 or 10 wherein the complex response adjuster comprises a complex-value multiplier.
25
        20.   The baseband processor of claim 3 or 18 wherein the amplitude adjuster comprises a real-value multiplier.

        21.   The baseband processor of claim 1 further
30   comprising a sample rate converter.

FIG.-1a

FIG.-1b

FIG. —2a

FIG. –2b
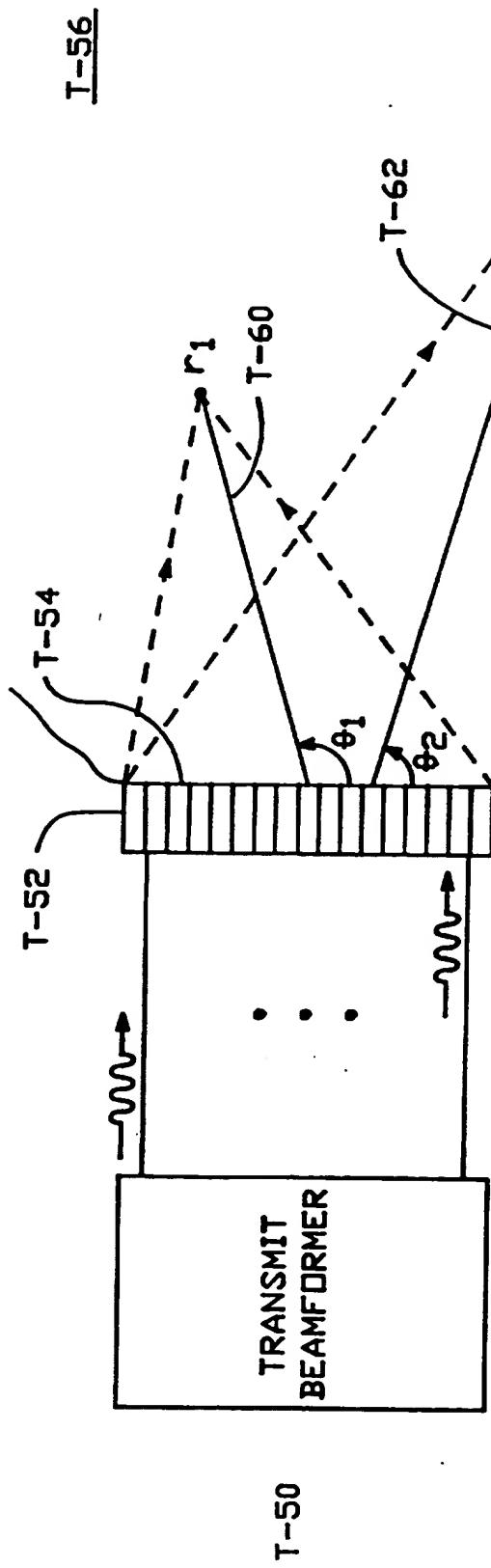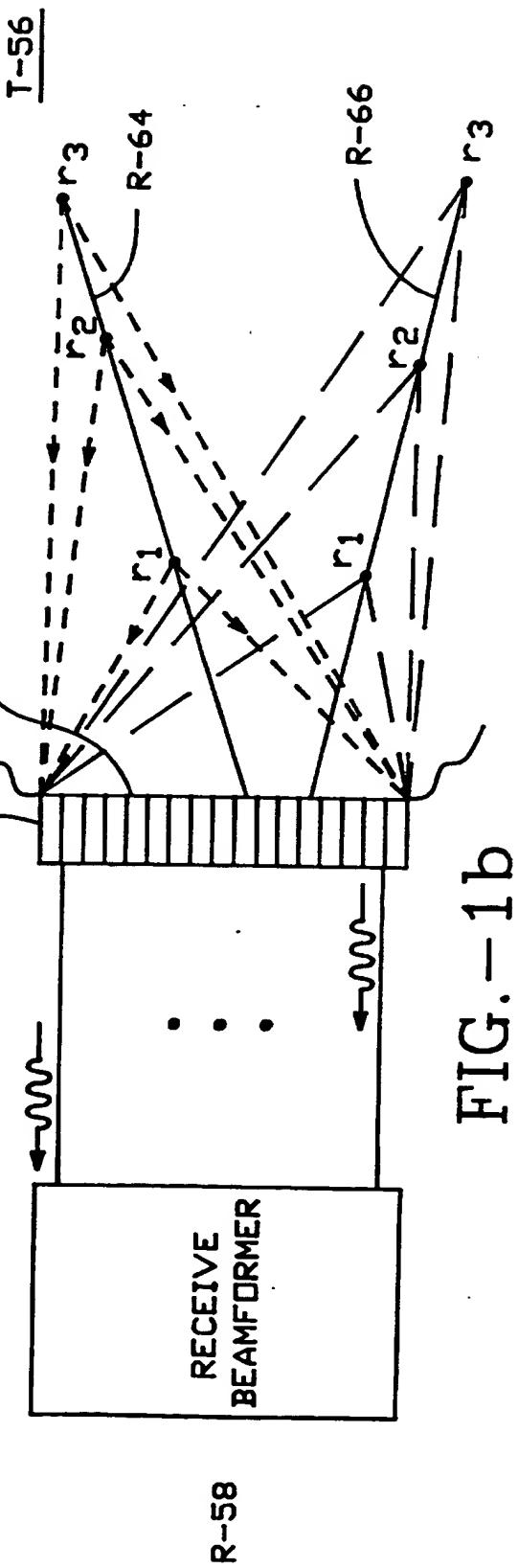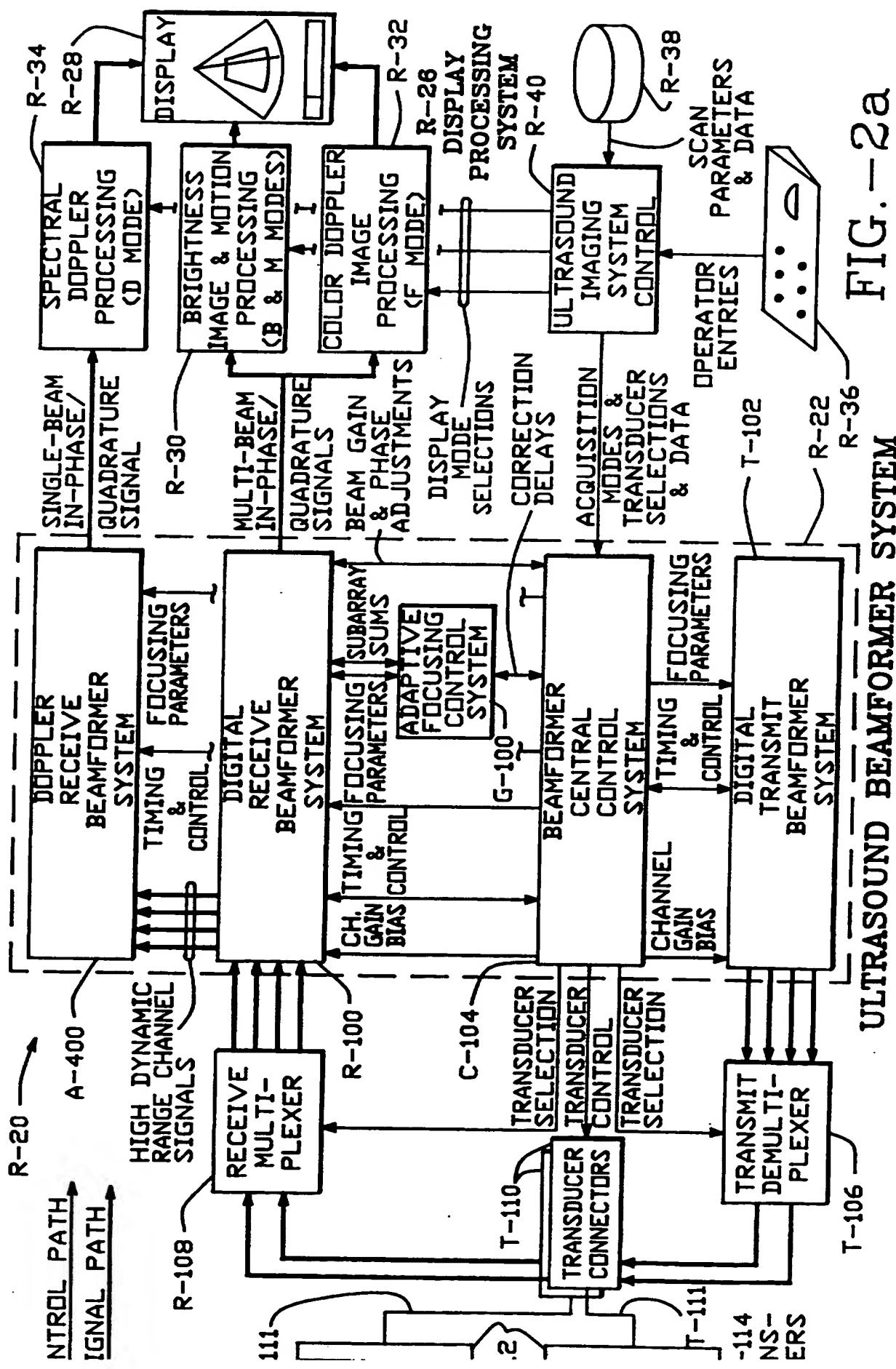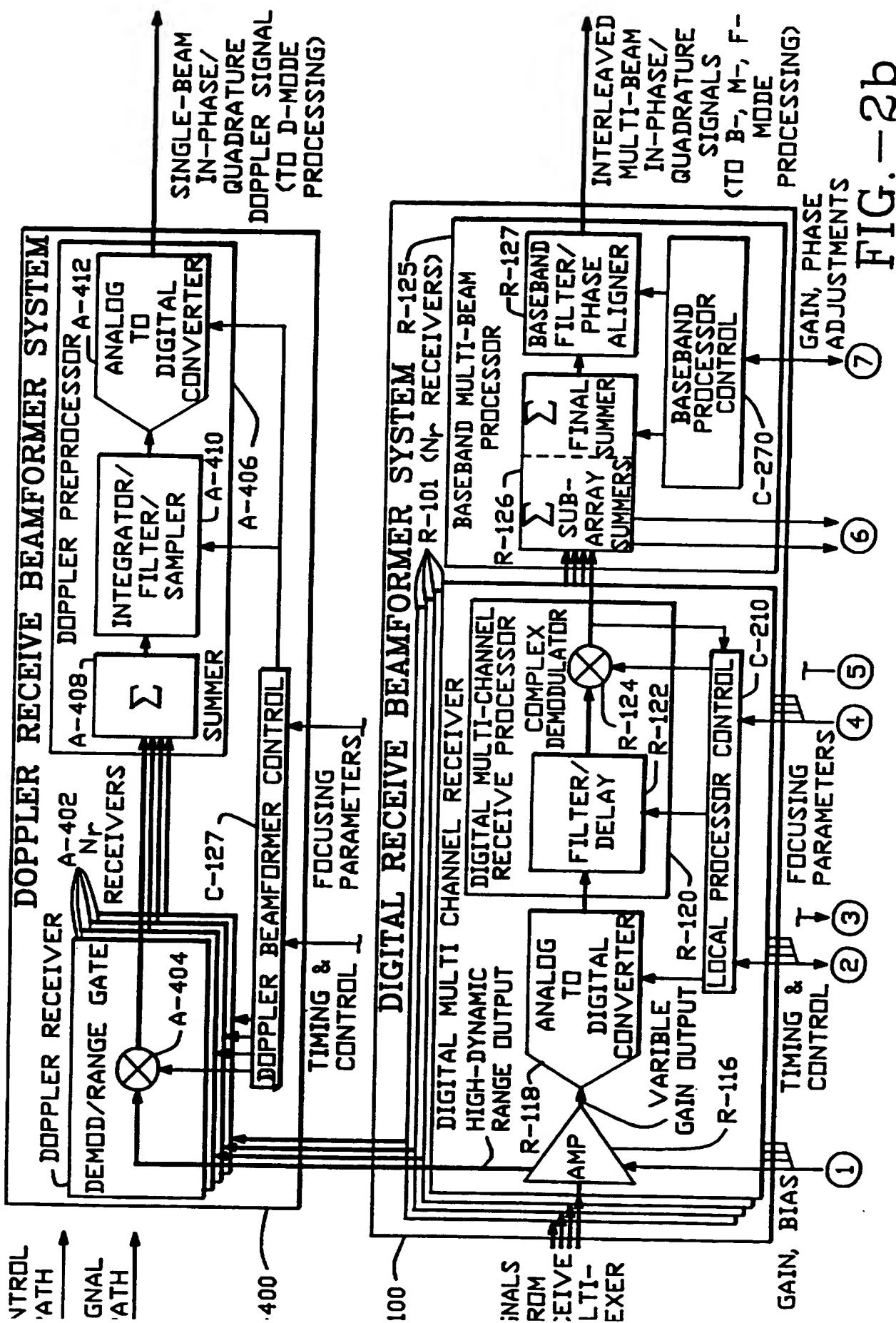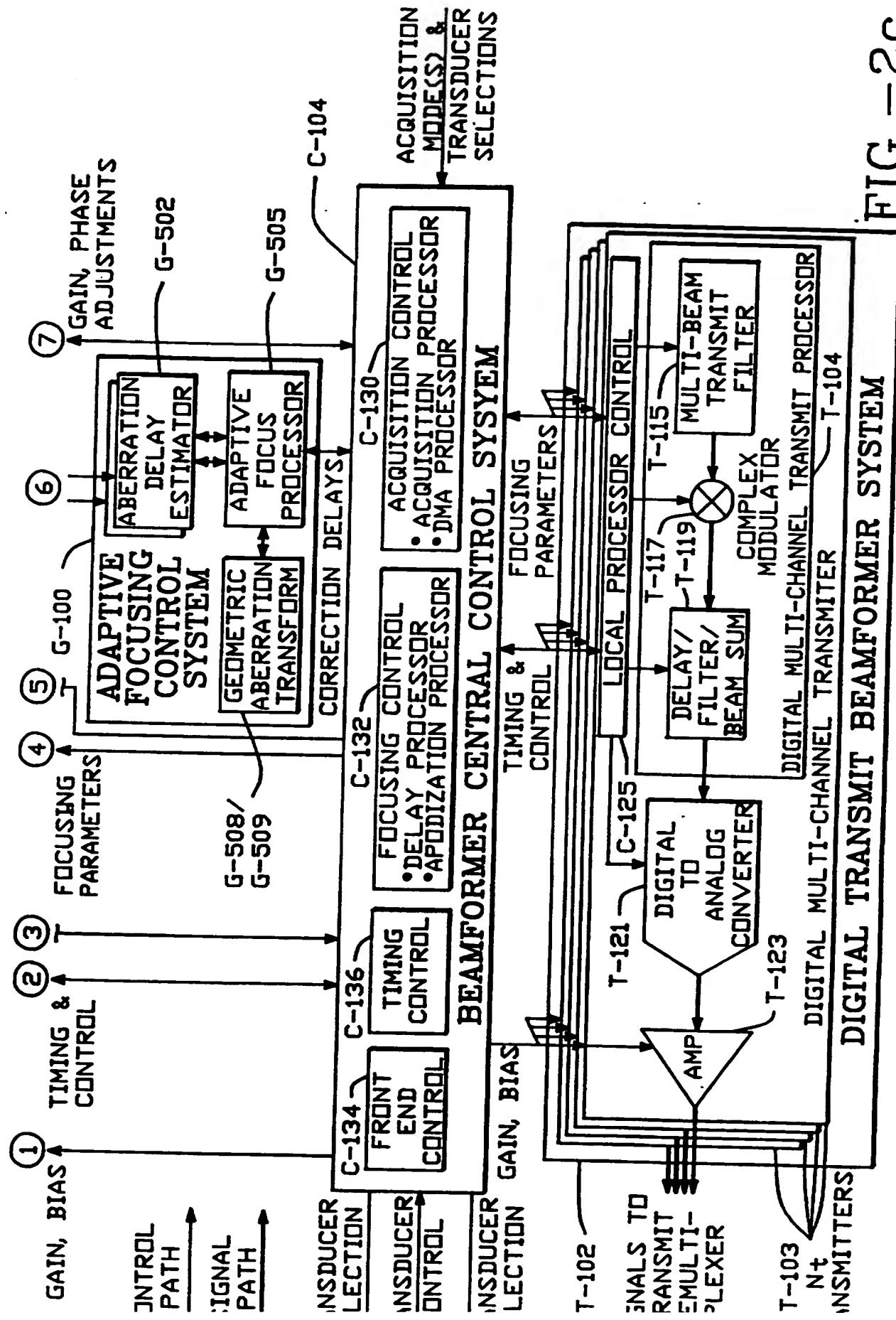
FIG.—2c
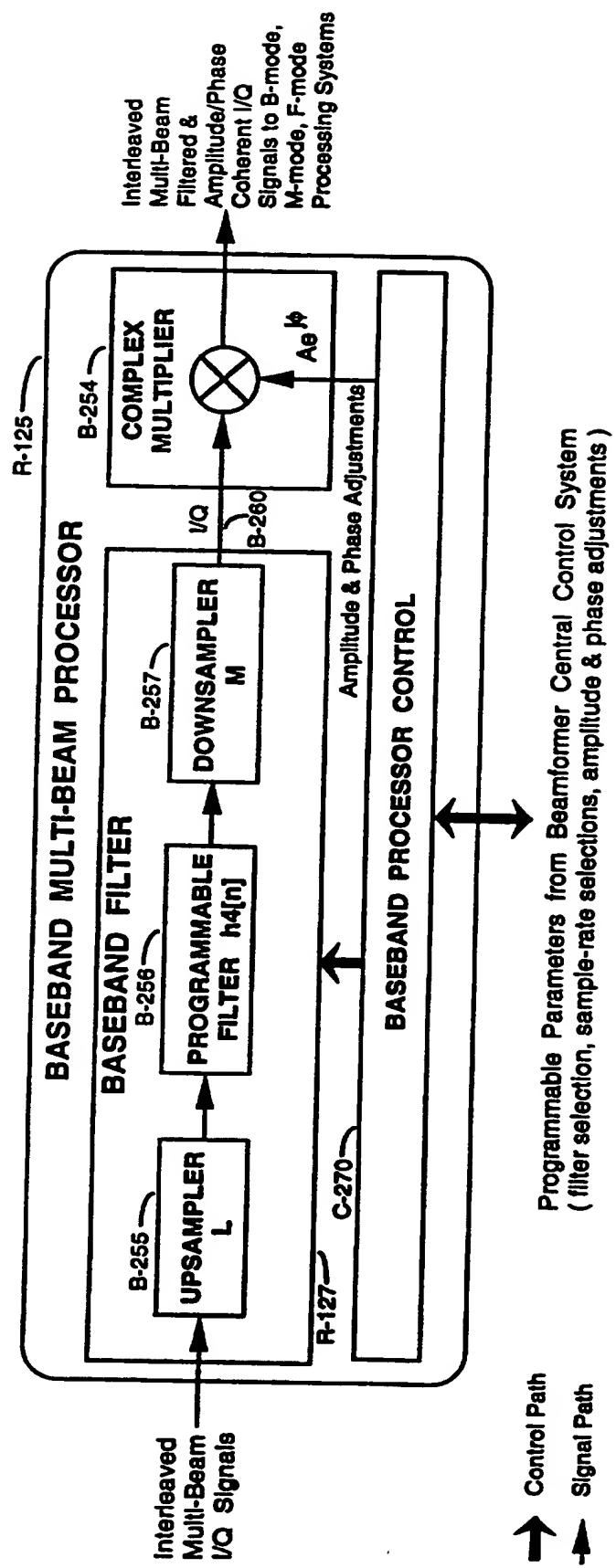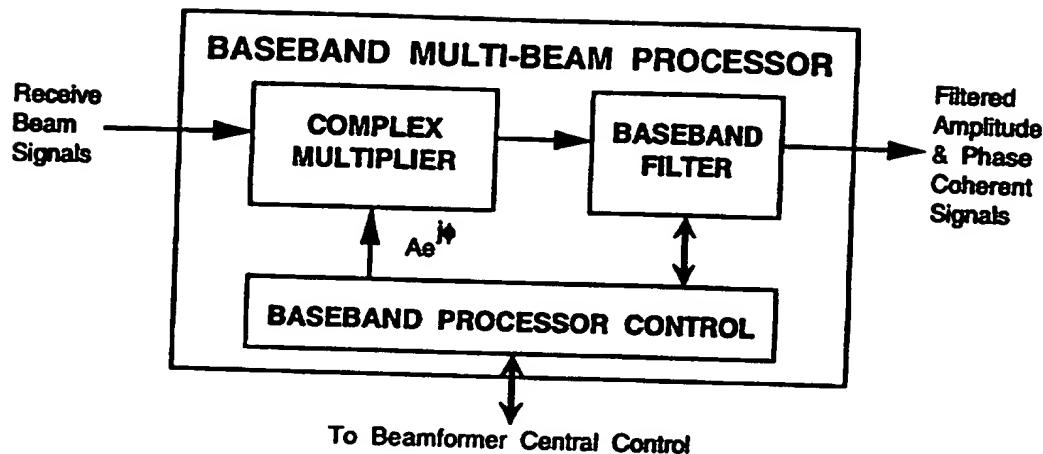
Figure 3.

Figure 4a


Figure 4b

FIG. 5

FREQUENCY PROFILE
VS
TIME

$f_P(t)$

F_{START}

t

FIG. 6a

$f_P(t)$

F_{START}

$F_{DOWNSLOPE}$

F_{LIMIT}

T_{BREAK}

t

FIG. 6b

$f_P(t)$

F_{START}

$F_{DOWNSLOPE}$

F_{LIMIT}

$F_{UPSLOPE}$

T_{BREAK}

t
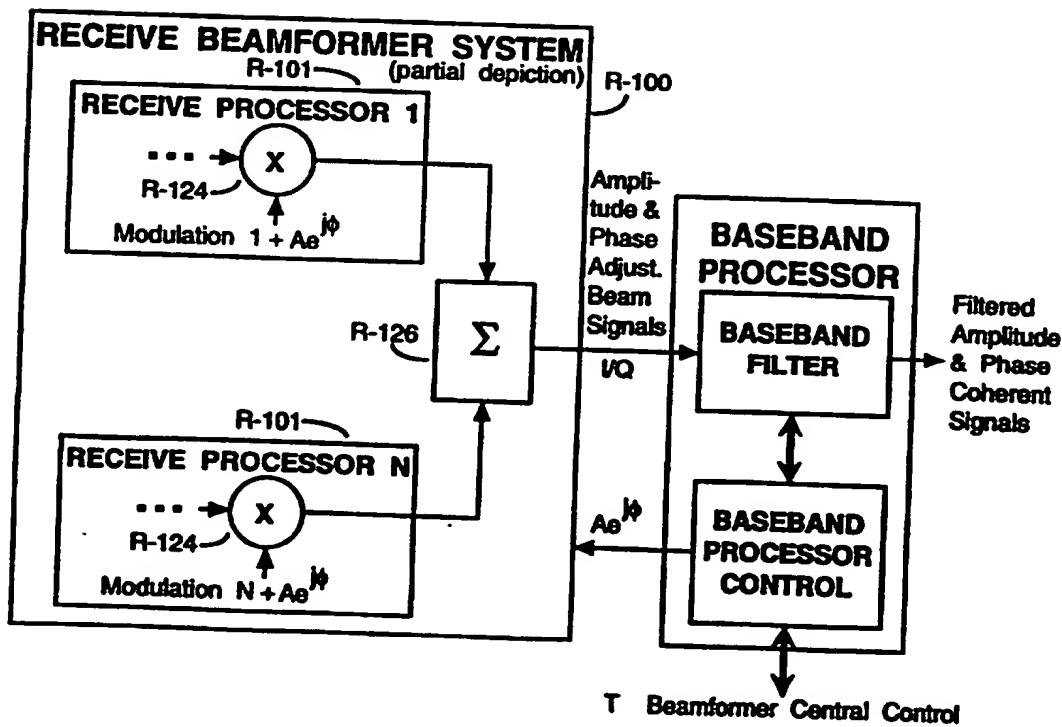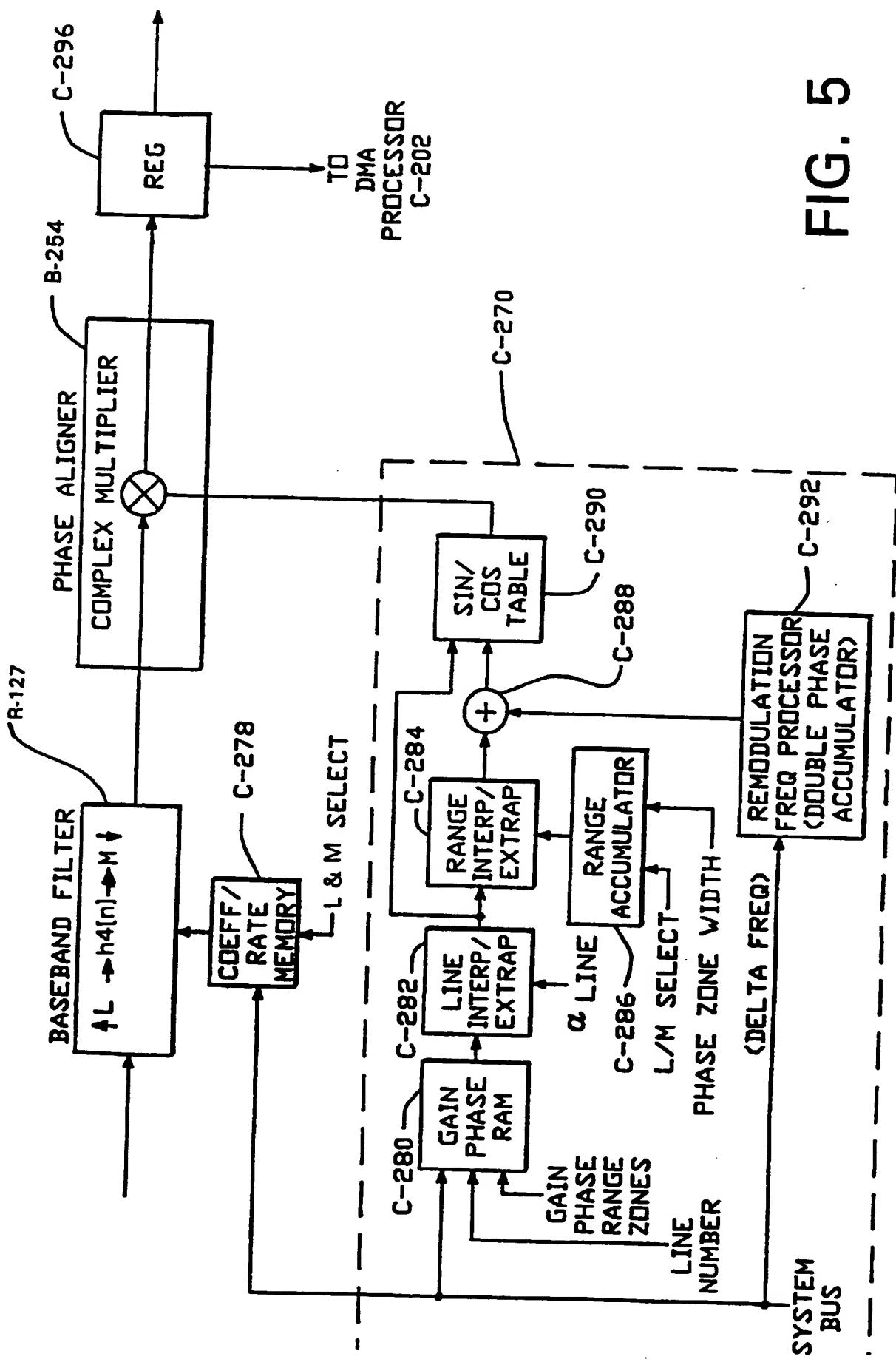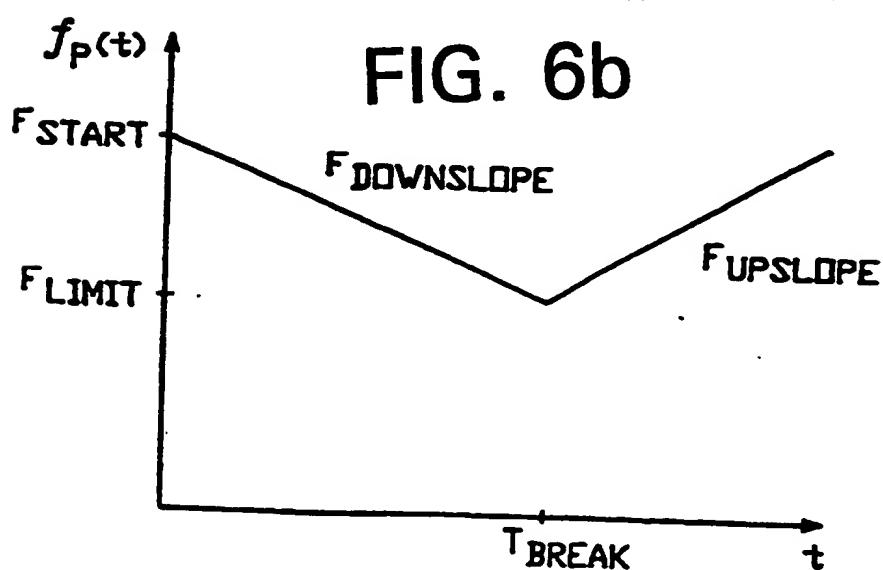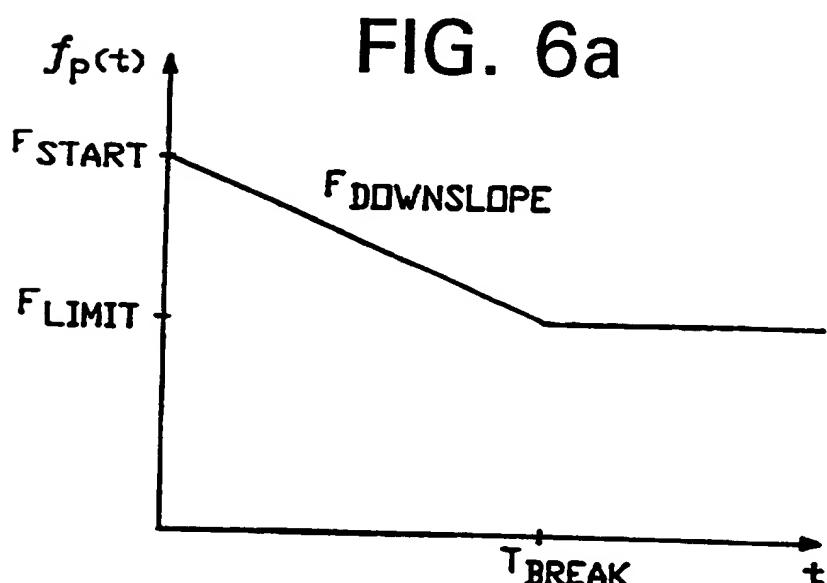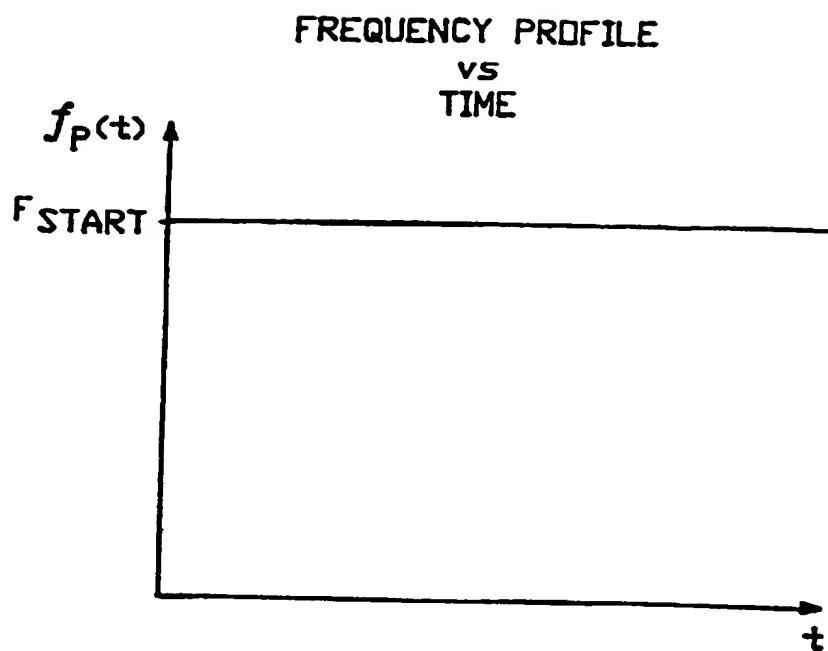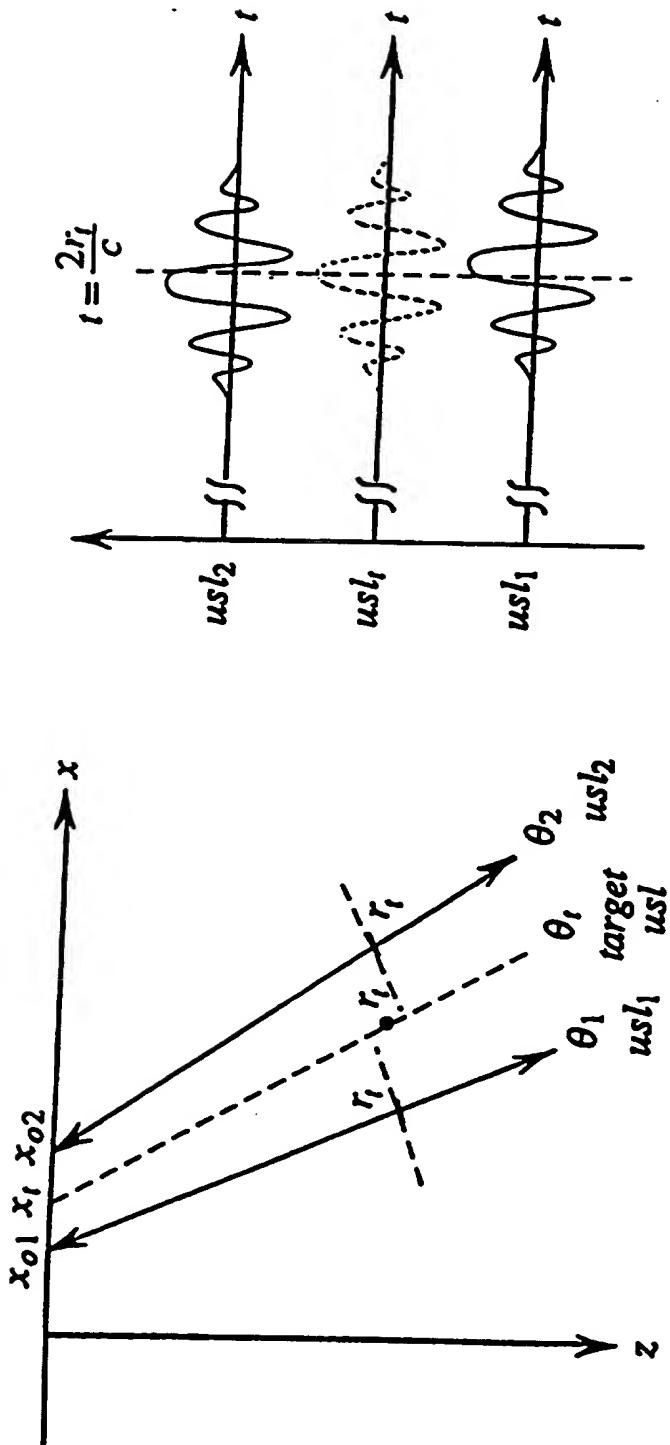
FIG. 7
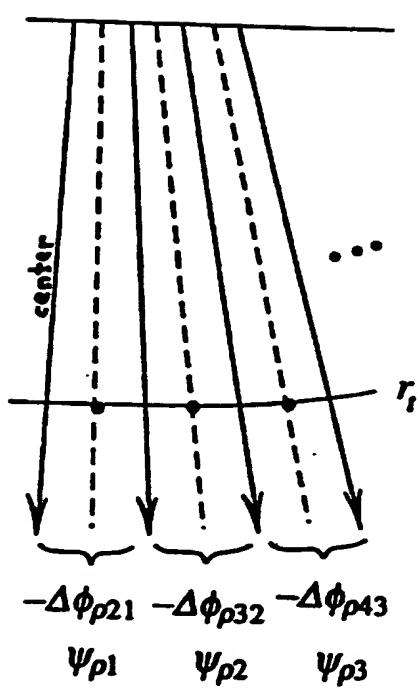
FIG. 8
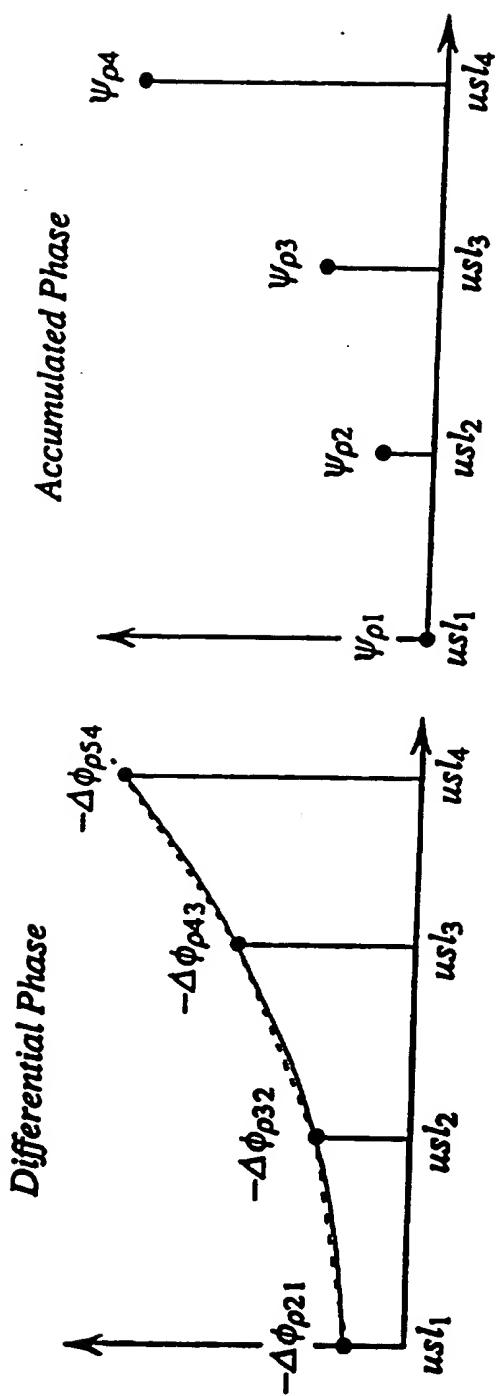
FIG. 9

International application No.

PCT/US95/09902

A.   CLASSIFICATION OF SUBJECT MATTER

IPC(6)   :A61B 8/00

US CL   :128/661.09

According to International Patent Classification (IPC) or to both national classification and IPC

B.   FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 73/861.25; 128/660.05, 661.08, 661.09; 342/160, 378, 379; 375/235, 344, 350, 375

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

NONE

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

NONE

C.   DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A	US, A, 5,170,792 (STURGILL ET AL.) 15 December 1992, see entire document.	1-21
A	US, A, 5,016,641 (SCHWARTZ) 21 May 1991, see entire document.	1-21

☐ Further documents are listed in the continuation of Box C.   ☐   See patent family annex.

*	Special categories of cited documents:	"T"	later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention
"A"	document defining the general state of the art which is not considered to be part of particular relevance		
"E"	earlier document published on or after the international filing date	"X"	document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone
"L"	document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)		
"O"	document referring to an oral disclosure, use, exhibition or other means	"Y"	document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art
"P"	document published prior to the international filing date but later than the priority date claimed	"&"	document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
04 OCTOBER 1995	06 NOV 1995

| Name and mailing address of the ISA/US Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231 | Authorized officer FOR GEORGE MANUEL |